Mixed carrier communication for sixth-generation networks:

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Mixed Carrier Communication for Sixth-Generation Networks

by

Ahmed F. Hussein

A Dissertation
Submitted to the University at Albany, State University of New York
In Partial Fulfillment of
the Requirements for the Degree of
Doctor of Philosophy

College of Engineering and Applied Sciences
Department of Electrical and Computer Engineering

2021
Declaration

I, Ahmed F. Hussein, declare that this thesis titled, “Mixed Carrier Communication for Sixth-Generation Networks”, and the work presented in it are my own. Studies discussed within the scope of this thesis have been either published or accepted in the following citations and I am the lead author of the works.


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Abstract

Recently, research on sixth-generation (6G) wireless networks has gained significant interest. By 2030, it is expected that 6G will introduce revolutionary applications and services. Thus, 6G is likely to expand across all available spectrum, including terahertz (THz) and optical frequency bands. Although 5G will offer a massive upgrade to the spectrum, the technology does not provide solutions to support a vast multitude of services and devices simultaneously.

Motivated by the heterogeneity of wireless technologies, devices, and services, the **Mixed Carrier Communication (MCC)** concept is introduced for the first time. MCC is a novel concept that supports the 6G vision by enabling simultaneous broadband access, low-rate Internet-of-Things (IoT) connectivity, device-free sensing, and device-based localization. Additionally, visible light communication (VLC) acts as a core technology for the introduced design motivated by 1. the unlimited and unregulated spectrum, 2. the green nature of the technology introducing substantial power savings to provide communication on top of illumination, 3. the availability of the indoor lighting infrastructure, and 4. the fact that indoor communications represent more than 80% of the communications capacity.

MCC enables a unified transmission physical layer (PHY) design to simultaneously connect devices with different complexities. Thus, the proposed approach addresses the huge spectrum needs by futuristic applications, as well as the multitude of services enabled by varying device capabilities. Additionally, MCC addresses the VLC system requirements of illumination constraints and dimming control that is an essential feature for lighting systems. Modeling, simulation, and experimental results highlight the possibility of capturing high-speed, low-speed, and localization information from the same transmitted MCC waveform based on the receiver design.
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Dedication

To the memory of my beloved father, Prof. Fahmy Hussein, my continuous source of inspiration. To my mother, Nagwa, and my sister, Ghada, who fully supported me to leave for my dream when they needed me. To my lovely family, Yasmine, Jolie, and Malek who joined me during this long journey and whose sacrificial care and love made it possible for me to complete this work.
Chapter 1

Introduction

1.1 Cellular Network Generations

In the 1980s, the first-generation (1G) of wireless cellular networks was introduced as analog voice services supporting a limited number of users. By the 1990s, the second-generation (2G), i.e. Global System for Mobile Communication (GSM), became a global standard that could provide primitive packet data services on top of voice calls. The third-generation partnership project (3GPP) in the 2000s proposed several Megabits per second (Mb/s) through high speed downlink packet access (HSDPA) and high speed packet access (HSPA). As the technology developed, the wide use of smartphones allowed new services including social networking and video gaming. Thus, new challenges of spectrum crunch, limited capacity, and latency started limiting the users’ experience. By 2010, the fourth-generation (4G) Long-Term Evolution (LTE) was introduced to promise high data rates, i.e. 300Mb/s, and low latency, i.e. 10ms [1]. By 2020, the fifth-generation (5G) proposes a new radio (NR) core network with colossal capacity based on enhancing spectral efficiency, offering spectrum up to 60GHz and employing densified deployments. Accordingly, 5G is mainly focused on serving,

- massive machine type communication (mMTC) that requires low cost and energy. This could benefit traffic monitoring or remote control of buildings.
• ultra-reliable and low latency communication (URLLC) that requires a high degree of reliabil-
ity and availability with minimal latency. This could be essential for vehicular communi-
cation for traffic safety or industrial automation.

• enhanced mobile broadband (eMBB) services that require high data rates.

1.2 Review on Visible Light Communication

In the past few decades, telecommunication traffic has been increasing dramatically, causing spec-
trum crunch. Optical communication has been introduced to complement rather than replace radio
frequency (RF) technologies. Optical communication can be described as data transmission using
light as the carrier signal in an unguided channel in the visible, infrared (IR), and ultraviolet (UV)
bands. The idea of using light as a carrier was first employed by the Greeks. In the fourth century
B.C., fire towers were used to transmit information between the capital and the Greek Empire dis-
tricts to report enemy attacks. In the 16\textsuperscript{th} and 17\textsuperscript{th} centuries, using fire signals at night and smoke
signals in daylight was a common method of communication. In 1792, the optical telegraph was
introduced by Chappe. The system used telescopes and human repeaters to connect 29 cities in
France using around 540 towers with an average coverage of 4830 km. The studies on electricity
resulted in the electrical telegraph, which replaced the Chappe system due to the need to use it only
in daylight and fine weather conditions [2][3].

In the 19\textsuperscript{th} century, the famous scientist Alexander Graham Bell invented the photophone. He
could manage to transmit his voice over a distance of 200m by modulating sunlight. Interest in
the photophone continued, where the German army developed one that used a tungsten filament
lamp with an IR transmitting filter as a light source in 1935. After the laser invention in the 1960s,
research has re-emerged to achieve some fundamental contributions in optoelectronic semiconduc-
tors, leading to breakthroughs in long-distance optical communication. In 1962, TV signals were
transmitted over a distance of 30 miles by MIT Lincoln Labs. In the 1990s, the research in com-
munication between ground and satellite became popular. The explosion of optical communication
applications, including all different forms of telecommunications, started in the 2000s. Due to the advancement of solid state lighting (SSL) and light emitting diodes (LEDs), visible light communication (VLC) has been emerging so fast. In 2003, VLC emerged to offer optical communication in the visible light spectrum, *i.e.* 400-750nm, for indoor applications. Since then, VLC research and standardization efforts have been going [4, 5]. For instance, the IEEE 802.15 wireless personal area network (WPAN) task group 7 has completed a physical layer (PHY) and a medium access control (MAC) standard for VLC under IEEE 802.15.7. Since 2018, the 802.11 task group bb is responsible for the development of a standard document for VLC [6].

The most discussed VLC systems offer wireless transmission while illuminating indoor spaces using existing illumination infrastructure [7]. These VLC systems have been proposed as an alternative indoor wireless access technology, and as a complementary wireless access technology that coexists with RF technologies such as the 802.11 WLAN 2.4GHz, 5GHz, and 60GHz. Coexisting with RF-based technologies could contribute to dealing with the RF spectrum crunch and coping with the exponential increase in traffic demand [8]. Moreover, VLC is considered an enabling technology for future Internet-of-Things (IoT) deployments [9].

1.2.1 Visible Light Communication Modulation Techniques

Due to the unique physical nature of light as a carrier, VLC modulation techniques have different requirements than RF. Unlike RF, a VLC carrier has no embedded information in the phase or frequency of light. VLC modulation employs IM at the transmitter and direct detection (DD) at the receiver [10]. In other words, the electrical signals are modulating the generated light intensity of a light source while the receiver, *i.e.* a photo-diode (PD), will capture the light variations and translate them into electrical signals at the receiver.

In the last few years, many research groups have tried to design different flavors of the intensity-modulation (IM) compatible single-carrier and multi-carrier modulating signals aiming for better power-efficiency, spectral-efficiency, and dimming integration. For instance, low-complexity single subcarrier VLC modulation schemes have been proposed, including on-off keying (OOK),
pulse width modulation (PWM), and pulse position modulation (PPM) \[11, 12, 7\]. Additionally, orthogonal frequency division multiplexing (OFDM) has been widely used for VLC systems to avoid inter-symbol interference (ISI) \[13, 14\]. Several research efforts have exploited OFDM-based links, where optimized performance could be obtained based on popular techniques such as adaptive bit-loading and power-allocation per subcarrier \[15\]. For example, high speed transmission using optical OFDM is demonstrated in an optimized lab environment to demonstrate the capabilities of OFDM in terms of adaptive power-allocation per subcarrier, adaptive bit-loading per subcarrier, as well as, analog/digital equalization techniques \[16, 17, 18\].

Different formats of VLC compatible OFDM techniques have been introduced in literature limited by the intensity modulation and direct detection (IM/DD) requirements of transmitting positive and real signals. In order to satisfy the positive and real constraints on an OFDM VLC signal, unipolar real modulation schemes are introduced. One of the well-known techniques that trades off power efficiency for the sake of higher spectral efficiency is DC-biased optical OFDM (DCO-OFDM). Another technique that trades off spectral efficiency and extended bandwidth for higher power efficiency is asymmetrically clipped optical OFDM (ACO-OFDM) \[19, 15, 20\]. Several techniques aim at achieving better spectral efficiency, such as hybrid ACO-OFDM (HACO-OFDM), layered ACO-OFDM (LACO-OFDM), enhanced ACO-OFDM (eACO-OFDM), enhanced unipolar OFDM (eU-OFDM), and spectral and energy-efficient OFDM (SEE-OFDM) \[21, 22, 23\].

Only a few works consider dimming control while maintaining reliable broadband communication links. From these works, reverse polarity optical (RPO)-OFDM has been presented as a dimming compatible technique for Gigabit VLC links \[24\]. Based on the same concept, dimming control is replaced by a parallel channel of low speed data using an OOK-like envelope in \[25\]. These systems commonly use low-cost discrete PDs or cameras on mobile devices to detect the transmitted light signals using the rolling shutter effect \[26\]. The concept of transmitting sinusoidal beaconing waveforms from individual LEDs for indoor sensing and gesture control is proposed in \[27\]. Recently, efforts are directed towards the network integration of VLC towards LiFi networks and coexistence with the wireless local area (WLAN), i.e. WiFi, networks \[28\].
1.2.2 Optical OFDM

Due to the non-linearity of the light sources and the dispersive nature of the VLC channel, ISI could be a challenge [29]. Thus, OFDM is introduced to VLC modulation to combat ISI and enable proper channel estimation and equalization. Using OFDM, VLC systems can offer higher data rates compared to serial modulation schemes, i.e. quadrature amplitude modulation (QAM). The complexity of frequency domain equalization, i.e. in OFDM, is less than the time-domain equalization, i.e. in single carrier modulation (SCM). Additionally, the system complexity is moved to the digital domain, where a phase or frequency correction in the digital domain is much easier than a complex design of an analog filter [30]. However, the main challenges of VLC-OFDM, i.e. optical OFDM (OOFDM), are the high peak-to-average-power ratio (PAPR) and the sensitivity to phase noise and frequency offset [31, 32]. In Fig. 1.1, a block diagram for a conventional radio frequency (RF) OFDM system is shown [13]. The data is generated and coded to enable forward error correction (FEC) at the receiver. The interleaver makes sure that the data is rearranged to avoid burst data errors due to channel imperfections. Usually QAM modulators are used to map the bits into complex symbols before being fed to the inverse fast Fourier transform (IFFT) block. The \(m^{th}\) time-domain sample of IFFT is

\[
x_m = \frac{1}{\sqrt{N}} \sum_{k=0}^{N-1} X_k exp\left(\frac{j2\pi km}{N}\right) \quad 0 \leq m \leq N - 1
\]  

(1.1)

where, \(k\) is the order of the subcarrier and \(N\) is the number of subcarriers, i.e. IFFT length. The cyclic prefix (CP) is utilized to replicate \(G\) samples from the end to the beginning of the

![Fig 1.1: A block diagram for a conventional RF OFDM system.](image-url)
time-domain vector [33][30]. Although this approach reduces bit rate by adding redundancy, it eliminates ISI and inter-carrier interference (ICI). The generated analog waveform is modulated using an in-phase/quadrature (IQ) modulator. Assuming no distortion at the transmitter side, the received signal after demodulation and removal of CP follows

\[ y_m = x_m + w_m \]  

where, \( w_m \) represents the receiver noise. Thus, the received constellations after fast Fourier transform (FFT) is

\[ Y_k = \frac{1}{\sqrt{N}} \sum_{m=0}^{N-1} y_m \exp\left(\frac{-j2\pi km}{N}\right) = X_k + W_k \quad 0 \leq m \leq N - 1 \]  

(1.3)

where \( W_k \) follows a gaussian distribution according to the central limit theorem [13]. Since the RF channel is a multipath fading channel, the receiver captures multiple versions of the transmitted signal with varying gains and delays. Thus, the transmitted subcarriers are impacted by a frequency selective channel and require equalization. Accordingly, the previous equation is modified to be

\[ Y_k = H_k X_k + W_k \]  

(1.4)

where, \( H_k \) represents the channel frequency response. Assuming a linear dispersive channel, where the received signal is a sum of two signals with different gains and delays, a single tap equalizer can reverse the impact of the channel [13][33] as follows,

\[ \hat{Y}_k = \frac{Y_k}{H_k} = X_k + \frac{W_k}{H_k} \]  

(1.5)

The requirements of a VLC-OFDM waveform are different from RF, where the generated waveform that feeds a light source has to be positive and real. Thus, a conventional RF OFDM is modified from generating bipolar and complex samples to generate positive and real samples to
suit the IM/DD nature of VLC signals. Accordingly, multiple versions of VLC-OFDM have been introduced to the literature with the most common ones, DCO-OFDM and ACO-OFDM.

In order to ensure a real output signal after the IFFT block at the transmitter, a Hermitian symmetry is employed according to [34, 18],

\[
X_k = X^*_N - k, \quad 0 \leq m \leq \frac{N}{2}
\]  

(1.6)

There are different ways to obtain a unipolar positive waveform. For DCO-OFDM, a DC-bias is added to the electrical signal before IM. This avoids negative peaks unless the signal has a high PAPR, resulting in negative peaks clipping and distortion. In ACO-OFDM case, all the odd subcarriers only are loaded with useful data while the even subcarriers are zeros. All negative values are clipped. The resulting clipping noise impact falls upon the even subcarriers. This approach results in time-domain antisymmetry as follows [18, 16, 34],

\[
x_m = -x_m + \frac{N}{2}
\]  

(1.7)

There is a clear trade-off between the power and spectral efficiencies based on DCO-OFDM and ACO-OFDM design. For DCO-OFDM, the added DC-bias increases power consumption and thus reduces power efficiency compared to ACO-OFDM. Due to the Hermitian symmetry, DCO-OFDM spectral efficiency is halved compared to conventional RF OFDM. ACO-OFDM spectral efficiency is half that of the DCO-OFDM as only the odd subcarriers are utilized. The normalized bandwidths for DCO-OFDM, i.e. \(B_{N(DCO)}\), and ACO-OFDM, i.e. \(B_{N(ACO)}\), for an M-QAM order are illustrated below [20].

\[
B_{N(DCO)} = \frac{\log_2 M}{1 + \frac{2}{N}}
\]  

(1.8)

\[
B_{N(ACO)} = \frac{\log_2 M}{2(1 + \frac{2}{N})}
\]  

(1.9)
Moreover, the optical channel characteristics are fixed for a given position of the transmitter, receiver, and intervening reflecting surfaces. In other words, with relatively slow motion of people and objects in an indoor environment, the channel varies only on the time scale of multiple bit periods. Thus, the optical channel is considered to be a quasi-static channel \[35,36\]. In general, an indoor optical channel is dominated by the line-of-sight (LOS) propagation model with minimal attenuation and scattering. The DC gain \( H(0) \) for an optical receiver located at distance \( d \) and angle \( \phi \) from the transmitter follows the Lambertian distribution and can be shown in (1.10) \[3\].

\[
H(0) = \begin{cases} 
\frac{A_r(m_1+1)}{2\pi d^2} \cos^{m_1}(\phi) T_s(\psi) g(\psi) \cos(\psi) & 0 \leq \psi \leq \Psi_c \\
0 & \text{otherwise} 
\end{cases} \tag{1.10}
\]

where, \( A_r \) is the detector active area, \( m_1 \) is the order of Lambertian emission, \( T_s(\psi) \) is the optical band-pass filter transmission, and \( g(\psi) \) is the imaging concentrator gain at the receiver side. Thus, the received optical power \( P_r \) at a distance \( d \) from a transmitter with an optical power of \( P_t \) follows,

\[
P_r = H(0)P_t \tag{1.11}
\]

### 1.2.3 ACO-OFDM Overview

In this section, the basic description of a conventional optical ACO-OFDM system is introduced, as shown in Fig. 1.2. The IFFT/FFT blocks on the transmitter and receiver sides, respectively, are the main functions that distinguish OFDM from single-carrier modulations. In RF systems, the input for IFFT is usually a vector of complex numbers representing QAM constellations. In this case, the output of IFFT is also complex. In an IM/DD baseband modulation as optical OFDM, complex samples need to be replaced by real-valued coefficients. To generate real valued samples after IFFT, the spectral efficiency is sacrificed based on the Hermitian symmetry concept. This means that the second \( \frac{N}{2} \) samples of the input vector \( X[k] \) to IFFT of length \( N \) have to be the
complex conjugates of the first \( \frac{N}{2} \) samples \([34]\). The IFFT input samples \( X[k] \) are expressed as

\[
X[k]_{k=0}^{N-1} = \begin{bmatrix}
0 & \{y[k]\}_{k=1}^{\frac{N}{2}-1} & \{y[k]^*\}_{k=\frac{N}{2}}^{N-1} & 0
\end{bmatrix}
\] (1.12)

where, \( \{X[k]^*\}_{k=\frac{N}{2}}^{N-1} \) are the complex conjugates of \( \{X[k]\}_{k=1}^{\frac{N}{2}-1} \).

**Figure 1.2:** Conventional ACO-OFDM block diagram illustrating the common blocks in the proposed system (i.e. in blue).

The output time vector from IFFT \( x[n] \) is fed into a parallel-to-serial (P/S) block, and then CP is added to prevent ICI. After that, the signal is converted into analog form utilizing a digital-to-analog converter (DAC). In RF systems, the analog signal is then modulated using an IQ modulator for up-conversion of the complex signal to the carrier frequency, which is not the case with optical baseband modulation. The Hermitian symmetry step ensures a real signal to be modulating the optical source in both DCO-OFDM and ACO-OFDM. Moreover, the light source that transmits the OFDM waveform and its driving circuit have a linear dynamic range that is preferred for communications. In this linear region, the relation between the output flux and the driving current is nearly linear. Consequently, an LED requires a driving DC-bias to operate in the linear region. This does not allow negative signal, i.e. driving current, values \([37]\). On the one hand, a DCO-OFDM signal is biased by a level proportional to the standard deviation of the electrical signal. Although this technically solves the problem of having negative values, DCO-OFDM is considered to be
less attractive compared to ACO-OFDM due to the high power requirements. On the other hand, in ACO-OFDM, only odd subcarriers are modulated, which creates a redundant anti-symmetric signal where, all data that exist on the negative values of the time-domain signal also exist on the positive values. Therefore, zero-clipping removes all the redundant negative values eliminating the need for extra DC-bias, i.e. above the level required for driving an LED, to ensure a positive signal. An ACO-OFDM signal before zero-clipping, showing the asymmetric shape, is presented in Fig.1.3.

\[ R_{ACO} = \left( \frac{N}{N + N_{cyc}} \right) \log_2 M \text{ bits/s/Hz} \quad (1.13) \]
where, $N$ is the IFFT length, $N_{cye}$ is the number of CP samples, $B$ is the bandwidth, and $M$ is the modulation order. On the receiver side, the optical signal is captured using a PD where, the signal is shifted to the electrical domain for processing. Then the analog signal is demodulated and converted into digital samples. The FFT process reattains the IQ samples, and then the bits are detected after a QAM demodulator. In this block diagram, some blocks that are used in standardized OFDM, i.e. for example, interleaving, coding, and single tap equalizer, are removed for simplicity and highlighting the proposed modifications on the system.

1.3 Research Motivation

After a decade from introducing 4G, this decade is considered to be the age of untethered cloud virtual reality (VR) headsets and always-connected wireless medical devices [38]. According to [39], the average annual data downloaded per user is expected to be 1 Terabyte by the end of 2020. The exponential rise in the number of users and traffic capacity has driven the efforts of introducing 5G to be standardized and soon commercialized. 5G is resolving the limitations of current wireless technologies that cannot meet the future demand of the high degree of heterogeneity in terms of services, device classes, deployment environments, and mobility levels [40]. Millimeter-wave (mm-wave) is set to be the new technology for 5G standards and to offer a huge spectrum that can support a data speed of 1Gbps at a round trip delay of around 1ms [41] [39].

5G is device-centric rather than base station-centric as in 4G, thus offers a new vision for the diversity of devices and associated services that can be supported. Therefore, 5G requires minimum latency, maximum reliability, and minimum human intervention to connect millions of sensors and smart grids. In addition, IoT 5G networks are designed to support the considerable bandwidth requirements of millions of devices to be embedded within smart environments and cities, including homes and transportation systems. Smartphones and tablets are to be complimented or replaced in some cases by wearable devices which require huge bandwidth for real-time monitoring of health
conditions and share them over the network for further processing and storage. However, 5G invests in more spectral and hardware resources rather than offering ground-breaking technology [42].

Research about the sixth-generation (6G) of wireless cellular networks is being conducted to serve communication demands by 2030. 6G is expected to solve the gaps in design that 5G will not fill in. 6G is visioned to be more revolutionary than any previous generation by offering novel infrastructure design, vast bandwidth, new PHY modulation techniques and enhanced secrecy, security, and privacy. 6G has stricter requirements, compared to 5G, such as Terabits per second (Tbps) data rates, ten times lower latency, and sub-centimeter geo-location accuracy [43]. Though mmWave is expected to provide Gbps data rates in 5G, Tbps are needed for 6G applications. Hence, different frequency bands are to be explored to support higher data rates and more coverage density, including Terahertz (THz) and optical frequency bands [44]. To support the strict 6G requirements and the newly introduced frequency bands, new air interface, and transmission technologies are required. Additionally, physical layer security and enhanced spectral efficiency become essential for connecting billions of sensors and devices within the IoT 6G network [45].

With 5G being ready for deployment and commercialization, a hype in data usage per user is foreseen. 5G is also expected to concurrently support a wide variety of services. However, the practical trade-offs associated with concurrent services require further investigations. 6G research is interested in resolving the poor mm-wave propagation performance within, from, and to indoor environments. Thus, RF spectrum and VLC are expected to be combined within 6G networks. VLC is expected to provide optical-fiber-like performance for 6G networks within short ranges, i.e. few meters. A new level of heterogeneity is added to wireless networks to support by design a variety of services offered by various capability devices on a heterogeneous network [46, 47].

Although 5G is ready to be out as a commercialized solution, there is still no enough research on how the variety of services will perform concurrently. In addition, with directed line-of-sight (LOS) mm-wave beams to be directed within smaller cell sizes, biological safety will be a concern.
for the vast majority of users. Therefore, 6G is expected to offer a revolutionary category of services by design. On top of that, additional services could be offered as indoor positioning which is currently still far from maturity, unlike the global positioning system (GPS) which is a reliable outdoor positioning system for most services. Up to 5G, base station heights are not considered a significant part of the system design where the wireless network is considered more of a two-dimensional nature. In 6G, heights are not neglected anymore. Thus, a 3D network is visioned to integrate terrestrial, airborne, satellite, water surface, and underwater networks to concurrently support high speed air crafts, unmanned aerial vehicles (UAVs) and underwater auto vehicles and divers, where underwater optical communication will play a significant role \cite{48,49,50}. Moreover, 6G networks are predicted to offer a new level of user immersion beyond virtual and augmented reality (AR) systems, where real-time virtual meetings can be conducted using holographic communication. In addition, tactile communication based on remote holographic interaction will allow reliable teleoperation and cooperative auto driving. Mobile devices are expected to be replaced by wearable devices and implanted sensors to offer human bond communication. This vision requires a novel network architecture including new PHY design, multiple access techniques, and heterogeneity of different spectrum bands \cite{51,46,52} to support this wide variety of heterogeneity in devices, technologies, and services while being constrained on transmitted powers, frequency bands, limited hardware complexity, energy efficiency and maintaining high levels of privacy and security \cite{48,53,54}.

Recent research discusses how LED-based VLC technology can be used for multiple services, including communication, positioning, and sensing. In addition to the dual functionality of VLC systems in providing simultaneous illumination and communication, addressing multi-service capabilities is necessary for VLC systems to mature into practical systems ready for deployment. In the presence of such multi-service VLC systems, for example, at retails, VLC enabled light fixtures could enable (1) beaoning signals for location and indoor navigation, (2) communication links with embedded systems in the illuminated areas such as retail labels and shelves, and (3) high speed Internet access for better customer experience. Currently, VLC prototypes do not offer more
than a service at once. In typical systems, the existing approach is to switch to different types of modulating signals, transmitters, and receivers when service changes. These workarounds however do not directly address the main problem by designing a communication technique that offers such a unique capability to VLC systems towards hardware and application-independent VLC deployments. Getting high-end and low-end receivers to capture and process the same waveform to reliably estimate the transmitted bits and identify the beaconing signal is a challenging research issue. Despite these different efforts, the literature does not offer a design framework to analyze such composite signals. In this work, a PHY design to support VLC is considered to efficiently support concurrent services essential to serve the needs of 6G networks.

One of the promising applications that MCC can be of great use is head mounted displays (HMDs) such as oculus rift, samsung gear, microsoft hololens, google glass, and HTC vive. The current AR/VR industry is promising for different fields, including gaming, industrial and military training, and medical research. Adopting the VLC technology within HMDs could combat some of the current technical challenges, such as the interference with ambient WiFi and bluetooth signals. Additionally, the LOS nature of VLC combats the RF challenges of signal attenuation and multipath fading. The importance of MCC for HMDs is that it can handle their different sensor types and data requirements. For instance, VR and AR headsets employ multiple sensors and cameras for head and eye tracking, obstacle detection, capturing posture information, and even facial expressions. Thus, The data rate requirements and the technology used for communication vary accordingly. MCC can offer a universal waveform that can simultaneously communicate with the varying needs of low and high data rates. Hence, MCC enables the concurrent connectivity of embedded signals used for high data rates to create VR/AR animations and low data rates for signaling, synchronization, and sensors diagnosis and control.

Similarly, MCC value arises in smart home and office environments. One use case scenario where localization is needed with control and communication is the adaptive network lighting control, where illumination adapts to daylight changes and user movements, and luminaires are controlled corresponding to a network of photosensors. Moreover, unmanned aerial vehicles (UAVs); being
used for multiple reasons including mobile access points, will benefit from MCC for simultaneous communication and autonomous homing and docking. Furthermore, MCC will enable the synchronous connectivity with a wide smart home network of IoT sensors with different data rate requirements including security cameras, smart switches, door locks, fire, humidity and proximity sensors. Thus, MCC can handle streaming data, metadata, control, and calibration signals all at once.

1.4 Thesis Contribution

In Fig. 1.4, a qualitative comparison between 5G and 6G communication explains how this thesis contributes to the research knowledge. Although 5G is offering a vast spectrum based on mm-Wave technology, 6G is expected to be offering more spectrum based on hybrid THz and optical systems. This heterogeneity in technologies offers plenty of bandwidth for signal transmission on top of more transmission reliability against channel imperfections. Meanwhile, classic encryption
methods to provide more secure networks are being challenged by increasingly powerful computers. Thus, PHY security technologies via VLC and heterogeneity in technologies become an effective solution and match with the 6G vision. Additionally, low energy consumption and long battery lives are key challenges for the 6G research. Thus, reduced complexity sensor designs are to be promoted to extend their battery lives without impacting their performance.

Accordingly, this thesis proposes, for the first time, novel concepts for futuristic PHY designs that support this 6G vision as follows,

- **Mixed Carrier Communication (MCC)** is introduced, for the first time, to enhance the spectral efficiency based on rethinking the PHY design. This work promotes a multi-service concept motivated by the heterogeneity of device capabilities, technologies, and services. The concept of MCC is proposed to serve the potential of service- and device-independent PHY communication that can support a diversity of concurrent services; without trading off the available spectrum. The simultaneous services that MCC can serve are broadband access, low-rate IoT connectivity, device-free sensing, and device-based localization. In addition, MCC is designed to minimize inter-service interference, to avoid impacting other services supported by MCC, including dimming control as an essential feature of commercial illumination systems. For the first time, to the best of the authors’ knowledge, this study:

- Explains MCC PHY design procedure, introducing possible design trade-offs and performance optimization.

- Investigates the spectral profile of MCC towards proper spectrum management.

- Studies MCC performance with surrounding sources of lighting and beaconing interference.

- Evaluates MCC performance based on theoretical models, MATLAB simulations, and an experimental proof of concept.
Inspired by the heterogeneity of services and devices offered by MCC for futuristic 6G networks, a conceptual vision and initial study for futuristic PHY designs to support energy efficiency and PHY layer security is introduced as follows,

- **Lightweight multi-carrier modulation for IoT** is proposed as a novel technique to support the reduced complexity of IoT devices. The VLC version of this approach is investigated within the scope of this thesis under the name of **Lightweight Optical Orthogonal Frequency Division Multiplexing (LwO-OFDM)**. This falls under the category of enhancing the energy efficiency and extending the lifetime of sensing devices. On top, it enables IoT sensors to perform tasks that were not previously a part of IoT sensing capabilities. For instance, the introduced concept enables simple IoT sensors to capture OFDM waveforms and obtain channel estimation information with minimum computational complexity. The same concept enables cooperative communication to add another tier of MCC, and thus more spectral efficiency is obtained.

- The **novel Wireless Link-Pairing (WiLP)** concept is proposed to support wireless network security and users’ privacy. Most research on hybrid networks has explored methods to increase the capacity as a key factor of increasing spectral efficiency. The novelty of this concept lies within introducing joint signal processing within different interfaces, *i.e.* technologies, to introduce a novel PHY security design that prevents an eavesdropper who does not have both VLC and RF interfaces from capturing and interpreting the received signal.

Both of the LWo-OFDM and WiLP concepts are at an early stage of the investigation. Thus, this study introduces these concepts, their preliminary investigation results, and a vision of how the advancement of these technologies falls within futuristic 6G networks. More details about the contribution within every concept are listed in the following chapters as per the thesis structure shown below.
1.5 Thesis Structure

This chapter has provided a brief introduction to the cellular generations, followed by a review of VLC modulation techniques. The research motivation and thesis structure are presented. The remaining chapters are organized as follows,

- Chapter 2 introduces the novel PHY VLC-MCC design, modeling, and simulation. A study for power constraints illustrates the design trade-offs.

- Chapter 3 proposes the spectrum management and interference analysis of MCC to evaluate the performance under different design parameters. Moreover, the transmitter and receiver designs are explained, including an arithmetic complexity analysis to highlight the validity of MCC reception by both high- and low-end receivers.

- Chapter 4 discusses MCC evaluation and simulation results. The performance trade-offs between different modulation techniques embedded within an MCC waveform are studied.

- Chapter 5 investigates hardware implementations and experimental evaluation of the introduced concepts.

- Chapter 6 highlights the futuristic heterogeneity concepts of LwO-OFDM for IoT connectivity and WiLP for PHY layer security. Theoretical modeling and evaluation of LwO-OFDM are introduced. Additionally, WiLP is evaluated in terms of eavesdropping immunity and bit-error-rate performance.

- Chapter 7 concludes the thesis and studies the future research plan to finalize the doctoral dissertation.
Chapter 2

Mixed Carrier Communication Design

2.1 Introduction

Currently, a portfolio of wide-/local-/personal-/body-area connectivity solutions based on standardized or proprietary technologies with different types of transceiver frontends, modulating waveforms, medium access protocols, and network features are configured to support a variety of connectivity, sensing, and IoT applications [55, 56, 57]. Debates on the benefits of SCM and multi-carrier modulation (MCM) schemes have led to the choice of different binary and $M$-ary schemes depending on the targeted application, aiming for better spectral efficiency, power efficiency, and more [58, 20, 29, 13]. For instance, current IEEE 802.11 WLAN, i.e., WiFi, standards are based on the choice of specific SCM and MCM schemes. All standards and efforts to evolve or revolutionize existing networks do not consider a holistic approach to simultaneously support multi-service communication and sensing for devices having different resource capabilities. Thus, a need for novel PHY design is essential to support the wide variety of services while maintaining a high quality of service while avoiding inter-service interference. MCC proposes a PHY design of a composite waveform that overlays SCM and MCM, i.e., OFDM, schemes with integrated beacons within the same device-independent waveform which can be processed by high-end and low-end devices, i.e., device-independent.

In the lighting industry, pulsed dimming is favored over analog dimming, where a PWM forward current is used to control the light source intensity in order to avoid color shifts [59]. Based on
the set duty-cycle of the pulse and the proper pulse duration $T_{PWM}$, an LED is driven to operate into “on” and “off” intervals. For example, the light is turned “on” for half the cycle to bring the LED to 50% brightness. The longer the “on” periods relative to the “off” periods, the brighter the LED becomes. The PWM-dimming relies upon the human eye’s ability to assimilate the overall average of the emitted light pulses. The rate of a few hundred to hundreds of thousands of pulses per second ensures that flickering is unnoticed [60].

Adopting an MCC waveform requires careful design steps to integrate multiple services, i.e. modulation techniques, simultaneously. The first version of MCC PHY, i.e. VLC-MCC, aims at generating a PWM-sampled layer of beacons, SCM, and MCM signals. Thus, it can enable positioning, sensing, and both low and high speed communication. Analog OFDM samples are conditioned to generate a PWM-like envelope in order to maintain the control of various dimming levels. The OFDM samples are responsible for high speed communications, while other IoT sensing and positioning services require less complex forms of communications with low data rate requirements. This is achieved by adding and manipulating beacon signals within the PWM-like envelope without trading off the available spectrum.

### 2.2 Beacon Generation and Analysis

Optical beacons are widely used for localization using optical sensors, i.e. cameras, acting as a low cost solution for robotics applications and personal mobile devices navigation in indoor environments [61]. Device-based localization relies on reference anchors to find the relative location of a device which is helpful for indoor navigation by end-users. However, a monitor-based localization system can be helpful in user tracking and getting anonymized information about activity patterns by end-users in an indoor space, such as shopping patterns inside a shopping mall which directs business owners to offer location-based services and increase their sales. For instance, accurate low cost VLC localization systems can be implemented based on the received signal strength indicator (RSSI) according to [62]. In recent work [63], an optical bidirectional
FIGURE 2.1: (A) Individual PWM sampled sinusoidal beacon, (B) A PWM sampled sinusoidal beacon followed by a PWM sampled MCM, i.e. OFDM waveform.

beacon transceiver is considered in a VLC indoor environment in order to mitigate optical shadowing. Moreover, beaconing is compatible with IoT indoor lighting systems where beacons are considered as identifiers (IDs) for smart devices with LEDs [64].

Beacons typically work in low frequency ranges to be compatible with low cost LED drivers. The light sources in a room can be assigned different encoded IDs based on patterns that ensure low cross correlation; however, a high refreshing rate requires short encoding lengths that will affect the cross correlation performance [65]. Another approach is to identify light sources by orthogonal frequency beacons. Although the beacon waveform can be a sinusoidal analog waveform, it is preferred to use a PWM-sampled beacon.
Additionally, PWM dimming control is preferred over analog dimming in the lighting industry due to the possible color temperature variation and the brightness non-linearity resulting from analog dimmers. However, a PWM-sampled beacon also carries PWM harmonics. In the analog domain, and using a low-pass filter (LPF), a simple passive RC-filter removes the non-desirable PWM harmonics and lets through the desired analog beacon. The Nyquist-Shannon sampling rate constrains the PWM sampling resolution, i.e. at least double the beacon frequency. The sampling resolution increases with the increase of the PWM frequency. The selection of sampling resolution versus beacon frequency is thus an application-dependent trade-off [66]. The amplitude of the beacon samples is represented by varying duty cycles. A duty cycle of 50% represents a beacon sample with an amplitude of half the peak-to-peak voltage, while 75% of the voltage can be obtained from a 75% duty cycle [67]. Thus, the modulating waveform is a series of pulses with a fixed amplitude ratio of the peak-to-peak voltage of the beacon waveform, i.e. $V_{pp}/k$, and varying duty cycles that are proportional to the sampling amplitudes of the analog sinusoidal signal as shown in Fig. 2.1A.

To get a better insight into the beacon encoding process, the modulated signal is investigated in both time and frequency domains. A PWM signal current $i_{PWM}(t)$ with a varying pulse width $T_H$ and a period duration $T_{PWM}$ can be expressed as,

$$i_{PWM}(t) = \begin{cases} 
I_H, & 0 \leq t < T_H \\
I_L, & T_H < t \leq T_{PWM}
\end{cases} \tag{2.1}$$

where $I_H$ is the high level “on-state” LED current and $I_L$ is the low level “off-state” LED current. The sampling duty cycle can be expressed as $d(n) = T_H(n)/T_{PWM}$ [68], where $n$ represents the beacon sample. The sampling modulated pulse width, $T_H(n)$, changes linearly with the instantaneous value of the beacon waveform, i.e. $b(n)$, as given by

$$T_H(n) = \frac{b(n) - b_{min}}{2 b_{min} f_b} \tag{2.2}$$
where \( f_b \) and \( b_{\text{min}} \) are, respectively, the frequency and the minimum peak amplitude of the beacon waveform. The PWM sampling waveform peak-to-peak voltage can be adjusted to be a fraction of the original analog beacon waveform, \( i.e. \ V_{pp}/k \), to utilize the full LED dynamic range. In order to mitigate flickering impacts on human eyes, the minimum PWM frequency \( f_{\text{PWM-min}} \) should be 200Hz. Hence, \( f_{\text{PWM}} \) has to be larger than double the beacon frequency \( f_b \) to achieve the Nyquist-Shannon sampling rate constraint.

Considering an MCM OFDM, beside beacon-to-PWM mapping, PWM can also be used to linearly represent the bipolar OFDM signal as shown in Fig. 2.1B [67]. This approach is valid for any SCM and MCM. It is worth mentioning that this approach can significantly contribute to addressing the issues of high power consumption that DACs require for high order SCM QAM and associated high PAPR in the case of OFDM [31]. In contrast to high PAPR associated with OFDM, the PAPR of such binary-level signals only depends on the symbol length. However, practically, this approach trades-off bandwidth efficiency for power efficiency. A better mapping scheme, \( i.e. \) Multi-level MCC, is introduced and evaluated based on theoretical models and results from practical measurements in real-world scenarios in the following sections.

### 2.3 Dimming Control

A PWM-like envelope is designed to support target illumination, \( i.e. \) dimming controlled, levels. The envelope consisting of two main parts, a PWM-sampled beacon and dimming control PWM duty cycles, is shown in Fig. 2.2. In the following sections, the process of conditioning high speed OFDM samples to form this PWM-like envelope will be explained. The dimming control duty cycles are adjusted based on the target illumination. As previously mentioned, \( f_{\text{PWM}} \) is constrained to ensure a flicker-free system while maintaining a sufficient beacon sampling rate. Moreover, \( f_b \) is designed to be in the low frequency range to avoid interfering with either PWM-like envelope harmonics or high speed OFDM subcarriers. Hence, \( T_{\text{PWM}} \leq 5ms \) to avoid human eyes capturing light flickering, while for highly accurate positioning applications the beacon refresh rate, \( i.e. \)
periodicity, is recommended to be as low as 20ms. For other applications that require positioning of a walking person in an indoor environment 100-350ms is an acceptable range for the refresh rate \[69\].

According to \[70\], a commercial phosphor-based white LED has a 3dB modulation bandwidth of less than 10MHz without hardware pre-equalization \[70\]. However, the proposed design can be extended to comply with faster systems of higher modulation bandwidths, \textit{i.e.} laser diodes (LDs), to offer Gigabit links. In addition, the typical number of OFDM subcarriers used in multiple wireless network standards such as IEEE WLAN 802.11 is 64 \[71\]. Using the same number of subcarriers, the subcarrier spacing becomes 156.25kHz. To mitigate PWM harmonics interference with OFDM subcarriers while maintaining flicker-free illumination, \(f_{PWM}\) is constrained according to,

\[
200Hz \leq f_{PWM} < 156.25kHz
\] (2.3)

and thus the beacon frequencies are constrained according to the Nyquist-Shannon sampling rate by,

\[
100Hz \leq f_b < 78.125kHz
\] (2.4)
The beacon duration within a PWM-like envelope and the refresh rate introduce a trade-off between the dimming control efficiency and the localization error. As the beacon periodicity increases, the localization error decreases while the dimming control levels become more constrained. This results from the PWM-sampled beacon offering 50% dimming regardless of its frequency due to the symmetry between positive and negative cycles of the sinusoidal beacon. Hence, the fixed duty cycles, \( d_f \), within the PWM-like envelope are adjusted based on the target full brightness ratio, FBR, beacon duration, \( T_b \), and beacon periodicity, \( T_r \), i.e. refresh time, as follows,

\[
d_f = \frac{1}{T_r - T_b} [T_r \cdot FBR - \frac{T_b}{2}]
\]  

(2.5)

The PWM-dimming relies upon the human eye’s ability to assimilate the overall average of the PWM-like envelope. Furthermore, the relation between human eye response to light and designed FBR is non-linear due to the eye pupil wideness at low brightness levels [24] and can be expressed as,

\[
Perceived\ Light\ (%) = 100 \times \sqrt{\frac{FBR\ (%)}{100}}
\]  

(2.6)

### 2.4 ACO-OFDM power constraints

Analog OFDM samples are conditioned to form the PWM-like envelope, representing the beacon and dimming cycles resembling PWM cycles for digital dimming control. Accordingly, signal-to-noise ratio (SNR) is independent of the brightness over a wide dimming range. This framework is first proposed in [72, 73], to serve the potential of a multi-tier waveform that can support different types of service in addition to dimming control and sensing. The high capacity stream of bits can be modulated by any of the unipolar optical OFDM techniques. Different formats of VLC-compatible OFDM techniques have been introduced in literature limited by IM/DD requirements of transmitting positive and real signals. For example, one of the well-known techniques that trade off power efficiency for higher spectral efficiency is DCO-OFDM. Another technique that trades off spectral efficiency and extended bandwidth for a higher power efficiency is ACO-OFDM [34].
This is shown in Fig. 2.3A where the OFDM samples, in blue, build the black PWM-like envelope for the beacon and the dimming control duty cycles. In this section, ACO-OFDM is analyzed to find the optimum power to avoid clipping effects on the high speed link performance. The real positive time domain digital samples representing ACO-OFDM can be represented by,

\[
x(n) = \frac{1}{\sqrt{(N)}} \sum_{k=0}^{N-1} X(k) \exp\left(j\frac{2\pi nk}{N}\right)
\]  

2.7

where, \( x(n) \) represents the time domain samples, \( X(k) \) represents the frequency domain samples, and \( N \) is the number of subcarriers, i.e. FFT/IFFT length. Only \( 1/4 \) of the subcarriers are utilized while maintaining a Hermitian symmetry and clipping negative samples to obtain a positive real waveform. ACO-OFDM pdf can be represented as a truncated positive Gaussian waveform as shown in (2.8), where \( u(x) \) is the step function [18]. This can be illustrated by Fig. 2.3B.

\[
f_X(x) = \frac{1}{\sqrt{(2\pi)\sigma}} \exp\left(-\frac{x^2}{2\sigma^2}\right)u(x) + \frac{1}{2}\delta(x)
\]  

2.8

Following the analysis in [74, 75, 20, 13], this section investigates the optimum ACO-OFDM power to mitigate clipping noise introduced by the PWM-like envelope and LED dynamic range i.e. non-linearities. This study assumes that the PWM-like envelope is adjusted to avoid clipping by the LED limited dynamic range, and that the waveform is optimally DC-biased around the linear range of the LED. Hence, ACO-OFDM inverted samples within the PWM-like positive cycles are only clipped if their power exceeds a specific threshold to be clipped by the lower PWM-like envelope level. The same applies to the non-inverted samples within the negative PWM-like envelope cycles. According to [74, 75], clipping an ACO-OFDM waveform of electric power \( P_{\text{elec}} \), results in a clipped signal, \( x_{\text{clip}}(n) \), with a truncated Gaussian probability density function (pdf). The clipped signal can be modeled as an attenuated signal with uncorrelated non-Gaussian clipping noise added on top [74]. The clipping noise, \( \text{Clip}(n) \), can be represented according to (2.9), where \( K \) represents an attenuation factor of the original signal and \( \Delta(n) \) represents the
Figure 2.3: (A) ACO-OFDM samples are conditioned to formulate a PWM-like encoded beacon within a PWM-like dimming control frame, (B) ACO-OFDM pdf with zero mean and unity variance.
clipped part of the signal.

\[ Clip(n) = (1 - K)x(n) - \Delta(n) \]  \hspace{1cm} (2.9)

According to (2.10), the attenuation factor, \( K \), can be estimated, assuming the top side of the signal only can be clipped, where, \( \lambda_{top} \) represents the ratio between optical clipping threshold, \( \eta_{top} \), and the ACO-OFDM signal standard deviation, \( \sigma \).

\[ K = \frac{1}{2} - Q(\lambda_{top}) \]  \hspace{1cm} (2.10)

The clipping noise has a clipping variance, \( \sigma^2_{clip} \), that follows,

\[ \sigma^2_{clip} = \frac{P_{elec}}{2}(1 - 4K^2) - \Delta\sigma^2 \]  \hspace{1cm} (2.11)

\[ \Delta\sigma^2 = \frac{P_{elec}}{2} - \sigma^2_{x_{clip}} \]  \hspace{1cm} (2.12)

where, \( P_{elec} \) represents the original signal electric power, \textit{i.e.} pre-clipping, and follows \( P_{elec}/2 = \sigma^2 \) due to the ACO-OFDM waveform antisymmetry and zero clipping. The transmitted optical power is obtained by getting the expected value of ACO-OFDM pdf from (2.8) to be \( P_{opt} = \sigma/\sqrt{2\pi} \). However, photo-detectors follow a square law, where the electric current is proportional to the optical power rather than the signal amplitude. According to the above analysis, the relation between the optical clipping ratio (\( \lambda_{top} \)) and \( \sigma^2_{clip} \) is shown by (2.13) and can be visualized in Fig. 2.4. This is useful to define the range of clipping noise and find optimum optical clipping ratio to mitigate clipping effects; hence, optimum ACO-OFDM transmitted power to minimize bit-error-rate (BER) is obtained.

\[ \sigma^2_{clip} = -2Q^2(\lambda_{top}) + Q(\lambda_{top}) + Q(\lambda_{top})\lambda^2_{top} - \phi(\lambda_{top})\lambda_{top} \]  \hspace{1cm} (2.13)
where, \( Q(u) \) represents the complementary cumulative density function (CCDF) and follows,

\[
Q(x) = \frac{1}{\sqrt{2\pi}} \int_x^\infty e^{-u^2/2} \, du
\]  

(2.14)

and,

\[
\phi(u) = \frac{1}{\sqrt{2\pi}} e^{-u^2/2}
\]  

(2.15)

As shown in Fig. 2.4, the clipping noise has more impact within a specific region of the optical clipping ratio, \( \lambda_{\text{top}} \). In other words, the transmitted power can be optimized to mitigate the clipping effects. Hence, BER is plotted for multiple modulation orders in order to find the optimum clipping ratio at which BER is minimum. According to [74, 75], ACO-OFDM BER can be approximated to follow,

\[
BER = \frac{4(\sqrt{M} - 1)}{\sqrt{M\log_2(M)}} Q\left(\sqrt{\frac{3\log_2(M)}{M - 1} \gamma_{\text{elec}}} \right) + \frac{4(\sqrt{M} - 2)}{\sqrt{M\log_2(M)}} Q\left(3\sqrt{\frac{3\log_2(M)}{M - 1} \gamma_{\text{elec}}} \right)
\]  

(2.16)
\[
\gamma_{\text{elec}} = \frac{2K^2}{\frac{\sigma_{\text{clip}}^2}{P_b(\text{elec})} + \frac{N_o}{E_b(\text{elec})}}
\] (2.17)

where, \( \gamma_{\text{elec}} \) represents the effective electrical SNR per bit, \( P_b(\text{elec}) \) and \( E_b(\text{elec}) \) are, respectively, the electrical power and energy per bit and \( N_o \) is the noise spectral density. In Fig. 2.5 BER is plotted for 8-QAM, 16-QAM, 32-QAM and 64-QAM over a range of \( \lambda_{\text{top}} \) between 1.6 and 3.5 to deduce
the optimum optical clipping ratio and thus, optimum transmitted power per modulation can be inferred. This study aims to design an adaptive technique that responds to channel degradation by optimizing the transmitted ACO-OFDM power to minimize BER and enhance the performance while maximizing spectral efficiency. The solid colored plots represent the estimated BER vs SNR based on (2.16) over incrementing steps of $\lambda_{top}$. As expected, the BER performance is enhanced as the SNR increases; however, increasing the signal power beyond a certain limit results in more clipping noise and thus, the performance gets worse. This is illustrated in every solid colored line where the BER decreases with increasing SNR but starts to degrade beyond the minimum BER peak, at which clipping impact starts to be noticeable. Furthermore, top solid colored curves represent fewer values for $\lambda_{top}$ with more clipping noise and thus, BER performance is worse than the curves below. As expected, BER performance gets worse as the modulation order increases.

The blue curve fit represents the optimum SNR to obtain minimum BER for every $\lambda_{top}$. To have a better insight, Fig. 2.6A shows the relation between optimum SNR and target BER for different modulation orders. In addition, Fig. 2.6B represents the optimum optical clipping ratio ($\lambda_{top}$) for the target BER. For instance, a target BER of $3.8 \times 10^{-3}$, i.e. the hard decision FEC limit, the recommended optimum values for SNR and $\lambda_{top}$ are summarized in the following table. For more clear design procedures, $\lambda_{top}$ is shown below in terms of electric power to deduce the optimum electrical power clipping threshold, $\eta_{elec}$, for a minimum BER performance. Hence, the PWM-like frame power is designed to minimize clipping noise impact and thus, BER degradation.

Table 2.1: Optimum values for SNR, $\lambda_{top}$ and $\lambda_{top-elec}$ to target FEC BER threshold of $3.8 \times 10^{-3}$.

<table>
<thead>
<tr>
<th>Modulation</th>
<th>SNR</th>
<th>$\lambda_{top}$</th>
<th>$\lambda_{top-elec}$</th>
</tr>
</thead>
<tbody>
<tr>
<td>8-QAM</td>
<td>16.7dB</td>
<td>1.899</td>
<td>4.76</td>
</tr>
<tr>
<td>16-QAM</td>
<td>19.6dB</td>
<td>2.1964</td>
<td>5.5056</td>
</tr>
<tr>
<td>32-QAM</td>
<td>22.3dB</td>
<td>2.4526</td>
<td>6.1478</td>
</tr>
<tr>
<td>64-QAM</td>
<td>25.1dB</td>
<td>2.6652</td>
<td>6.6807</td>
</tr>
</tbody>
</table>

$$
\lambda_{top-elec} = \frac{\eta_{elec}}{P_{elec}} = \lambda_{top}\sqrt{2\pi}
$$

(2.18)
Figure 2.6: Curve fit for BER vs (A) SNR and (B) optimum values for $\lambda_{top}$ deduced from Fig. 2.5 for different modulation orders.

2.5 Beacon Modulation

In this section, beacon modulation design is proposed to explain overlaying SCM over MCM and beaconing within a PWM-like envelope. Thus, a low capacity link can serve wireless IoT connectivity. Inspired by SCM phase-shift keying (PSK) and PPM, a transmission layer could be realized by controlling the phase and the beacon position within the frame. In other words, both position and phase of PWM sampled beacons can be manipulated to modulate low capacity bits for IoT applications using low-end devices. PWM cycles are rearranged based on the target phase to modulate a stream of bits called beacon phase shift keying (BnPSK).

As shown in Fig. 2.7, a beacon is manipulated to represent different phases that correspond to modulated bits based on quadrature phase-shift keying (QPSK) modulation. As a result, demodulating the beacons in Figs. 2.7A-2.7D results in a bit pattern of “00”, “01”, “10” or “11”. In Fig. 2.7E another layer of SCM is added by manipulating the position of the beacon within the full frame. This beacon-position modulation (BPM) is inspired by the PPM concept. In the figure, 8-BPM is illustrated for N frames, while every individual frame can be controlled to provide a different brightness level.
Figure 2.7: A PWM encoded beacon showing corresponding duty cycles to beacon samples. In (A) to (D), beacon’s phase control from 0° to 270° is shown to represent Quadrature BnPSK modulation as an example of convenient modulation techniques to low-end receivers, and (E) BPM illustration with 8 different time-slots per frame, as an example, to modulate the transmitted bits based on BPM concept.
2.6 Summary

In this section, MCC is introduced as the first device-independent transmission that provides concurrent wireless services including broadband access, low-rate IoT connectivity, device-free sensing, and device-based localization. The design procedure of MCC is explained. First, the PWM-like envelope is generated to provide the required illumination level. A beacon is PWM-sampled and embedded within the PWM-like envelope to identify every light source using a different beacon frequency. The beacon phase and position within a PWM-like envelope are varied to provide transmission for low-end devices using BPM and BnPSK. Overlaying OFDM samples to provide MCM within MCC is constrained by the PWM-like envelope dynamic range. The OFDM power constraints are discussed with recommended SNR thresholds for every modulation technique to avoid performance degradation. For instance, an optimum SNR threshold of 25.1dB is recommended for 64-QAM to avoid clipping impacts.
Chapter 3

MCC Waveform Generation and Detection

3.1 Introduction

In this section, details about the generation and the detection of the designed MCC waveform are provided. Additionally, a spectral profile analysis of the MCC waveform is discussed to mitigate possible harmonic interference between different embedded SCM and MCM modulation techniques within the MCC envelope. The transmitter and receiver designs are discussed. A complexity analysis of the proposed receivers illustrates the possibility of detecting the MCC waveform using both high-end and low-end receivers. This gives an insight into how valuable MCC can be for IoT receivers with limited complexity and power capabilities.

In Algorithm 1 details of the MCC envelope structure describe the generation and detection process. A set of parameters are defined including the PWM frequency, $f_{\text{PWM}}$, the beacon frequency $f_b$ and the target FBR. The beacons are then PWM encoded by converting the beacon sample amplitudes to PWM-varying duty cycles. A phase shift is added based on the presence of BnPSK modulated bits. The PWM fixed duty cycle, $d_f$, for the dimming control part of the frame is decided based on the target FBR. The beacon position within the frame is controlled based on the BPM information bits. The entire MCC envelope is generated with OFDM samples conditioned to resemble the designed PWM-like envelope.

The detection process is illustrated in Algorithm 2. The low speed bits and localization IDs detection is separated from the high speed bits detection. Thus, both high-end and low-end receiver
**Input:** OFDM bits

Low capacity bits

**Output:** MCC waveform $x_{MCC}(t)$

**Initialization:**

- Define beacon frequency $f_b$
- Define PWM frequency $f_{pwm}$
- Set frame duration $T_r$ // beacon refresh rate
- Set beacon duration $T_B$ // Includes one or more beacon cycles
- Set target frame brightness ratio FBR
- Define OFDM parameters // FFT length $N_{FFT}$, Sampling Frequency $f_s$

**Frame generation:**

1. $spb = \frac{f_{pwm}}{f_b}$ // Number of PWM samples per beacon
2. $Deg_{stp} = \frac{2\pi}{spb}$
3. if PSK Modulation then
   4. Calculate phaseshift;
   5. else
   6. phaseshift = 0;
   7. end
8. $Deg = \text{phaseshift : } Deg_{stp}: \text{phaseshift} + 2\pi$
9. $d_B = (\sin(Deg) + 1)/2$ // Calculate beacon duty cycles
10. Generate PWM sampled beacon with varying duty cycles ($d_B$)
11. $d_f = \frac{1}{T_r - \frac{T_B}{2}} [T_r - FBR - \frac{T_B}{2}]$ // Calculate dimming responsible duty cycles
12. Aggregate beacon and dimming frame
13. if BPM Modulation then
   14. Locate beacon within frame based on BPM bits;
   15. else
   16. Locate beacon at the beginning of the frame;
   17. end
18. Generate ACO-OFDM samples
19. OFDM samples formulate a PWM-like frame and generate $x_{MCC}(t)$ // Aggregated beacon and dimming frame

---

**Algorithm 1:** MCC waveform generation algorithm.

Types can capture the same MCC waveform and obtain the relevant information. The low speed receiver utilizes a low speed PD or a LPF circuit to filter out all unwanted high speed harmonics. An FFT block can detect the ID for every light source where triangulation can locate an object with respect to multiple light sources. The beacon phase and position within an envelope can be demodulated into low speed bits. However, a conventional OFDM receiver can still capture the
Algorithm 2: Received MCC waveform detection algorithm.

high speed bits that are embedded within the same MCC waveform.

3.2 Spectrum Analysis and Interference Management

Due to the nature of the proposed MCC waveform, a thorough spectrum management procedure must be in process to avoid any possible inter-service interference. Improper spectrum management among the different services and lack of interference analysis leads to the degradation of the received signal quality. In this section, an analytical framework is presented to investigate the spectral synthesis of MCC waveforms. The composite MCC waveform is genuinely analyzed from a novel perspective compared to the existing approach in the literature, where analytical investigation of baseband ACO-OFDM is considered and followed by the spectral analysis of the composite waveform. Proper spectrum management is controlled by the design key parameters such as the number of OFDM symbols per PWM-like cycle, the PWM-like duty cycle and the frequency components of the PWM-like envelope as well as the OFDM subcarriers.

To understand the design key parameters affecting the performance of a composite MCC waveform, the following equations represent the time-domain analog waveform shape and spectral
profile. In (3.1), $x_h(t)$ represents the $h^{th}$ time-domain baseband ACO-OFDM symbol within a frame of OFDM symbols based on [30] and [76]. In (3.2), a PWM-like envelope of consecutive 50% duty cycles, $E_{PWM}(t)$, represents the dimming control envelope of the composite waveform. The 50% duty cycle is initially chosen to demonstrate the analysis in a simplified form before being generalized for varying duty cycles. The samples of $H$ OFDM symbols are reformed to take the PWM-like envelope shape represented in (3.2). The samples of the OFDM symbols are transmitted consecutively within this PWM-like envelope, where the samples that form the odd PWM-like cycles are DC-biased with amplitude $A$ after being reversed, while the samples that form the even PWM-like cycles are kept as is. This is detailed in (3.3) which results in the generation of the dimming control time-domain composite waveform, $w_d(t)$, as shown in (3.4).

$$x_h(t) = \sum_{k=-\frac{N}{2}}^{\frac{N}{2}-1} a_{k,h} e^{j2\pi f_k(t - hT_{OFDM})} \quad (3.1)$$

$$E_{PWM}(t) = \sum_{h=1}^{H} \text{rect}\left(\frac{t - (2h - 1)\frac{T_{OFDM}}{2}}{T_{OFDM}}\right)(-1)^h \quad (3.2)$$

$$E_{DC}(t) = \sum_{h=1,3,5,\ldots}^{H} A\text{rect}\left(\frac{t - (2h - 1)\frac{T_{OFDM}}{2}}{T_{OFDM}}\right) \quad (3.3)$$

$$w_d(t) = [E_{PWM}(t) \times \sum_{h=1}^{H} x_h(t)] + E_{DC}(t) \quad (3.4)$$

where, $N$ is the length of the FFT/IFFT, i.e. number of subcarriers, $a_{k,h}$ is the data symbol modulating the $k^{th}$ subcarrier in the $h^{th}$ OFDM symbol period, $f_k$ is the frequency of the $k^{th}$ subcarrier and $T_{OFDM}$ is the OFDM symbol duration. In the previous set of equations, the generation process of the composite waveform in time-domain is clearly shown. The same set of equations are used to synthesize the spectral profile of the designed waveform. The following set of equations clarify how the signal is presented in the frequency-domain. In (3.5), $X_h(f)$ represents the spectrum of a single OFDM time symbol. In (3.6), $W_d(f)$ represents the spectrum of the 50% dimming control
FIGURE 3.1: (A) Spectrum of a 50% MCC waveform with overlapping PWM-like harmonics and odd OFDM subcarriers, i.e. $C_2 = 3$, while (B) shows the spectrum of a waveform with non-overlapping PWM-like harmonics and OFDM subcarriers at different duty cycles, i.e. $C_2 = 10$.

composite waveform, $w_d(t)$, previously shown in (3.4). This set of equations highlights the spectral profile of a composite waveform with a 50% PWM-like envelope of one OFDM symbol per PWM-like cycle. However, as previously explained the practical scenario requires more flexibility over the duty cycles for dimming control, as well as the number of OFDM symbols per PWM-like
cycle to obtain higher bandwidth.

\[ X_h(f) = \sum_{k=-N}^{N-1} \frac{a_{k,h} T_{\text{OFDM}}}{\sqrt{N}} (-1)^k \frac{\sin(\pi f T_{\text{OFDM}} - k)}{\pi (f T_{\text{OFDM}} - k)} \]  

(3.5)

\[ W_d(f) = \frac{1}{H} \left[ \sum_{h=1}^{H} X_h(f) \exp(-j \phi_h)(-1)^h + A \frac{\sin(\pi f T_{\text{OFDM}})}{\pi f T_{\text{OFDM}}} \sum_{h=1,3,5,\ldots}^{i} \exp(-j \phi_h) \right] \]  

(3.6)

where, \( \phi_h = 2\pi f(2h - 1) \frac{T_{\text{OFDM}}}{2} \). In (3.7), \( W_{\text{dim}}(f) \) illustrates the spectrum of an RPO-OFDM signal with more generalized design parameters, for instance, varying PWM-like duty cycles, \( D \), and varying number of OFDM symbols per PWM-like period, \( T_{\text{PWM}} \).

\[ W_{\text{dim}}(f) = W_{\text{dim1}}(f) + W_{\text{dim2}}(f) \]  

(3.7)

\[ W_{\text{dim1}}(f) = \sum_{l=1}^{L} \sum_{h=1}^{C_1} X_{l,h}(f) \exp(-j [\phi_h + \phi_l + \pi]) - \sum_{h=C_1+1}^{C_2} X_{l,h}(f) \exp(-j [\phi_h + \phi_l + \pi]) \]  

(3.8)

\[ W_{\text{dim2}}(f) = \frac{A C_1 T_{\text{OFDM}}}{L T_{\text{PWM}}} \frac{\sin(\pi f C_1 T_{\text{OFDM}})}{\pi f C_1 T_{\text{OFDM}}} \exp(-j [\phi_h C_1 2h - 1 + \phi_l]) \]  

(3.9)

where, \( \phi_l = 2\pi f(l - 1)T_{\text{PWM}} \), \( L \) is the number of PWM cycles per frame, \( X_{l,h}(f) \) represents the spectrum of the \( h^{th} \) OFDM symbol in the \( l^{th} \) PWM cycle in the frame, the number of OFDM symbols per \( T_{\text{PWM}} \) is represented by \( C_2 = T_{\text{PWM}}/T_{\text{OFDM}} \) and \( C_1 = D C_2 \).

In order to have a better insight on how the previous analysis could help in the signal synthesis, MCC time domain waveforms and frequency domain profiles are shown when \( C_2 \) is equal to 2, \( i.e. \) 2 OFDM symbols per PWM-like cycle, in Figs. 3.2A and 3.2C and when \( C_2 \) is equal to 10 in Figs. 3.2B and 3.2D. The signal conditioning process differs in both cases. For instance, in the former case, each OFDM symbol with duration, \( i.e. T_{\text{OFDM}} \), of 1\,us fits \( T_{\text{PWM}} \) of 2\,us. The positive and negative samples are conditioned to shape a PWM-like envelope of consecutive positive and negative cycles. In this case, \( C_2 = 2 \), the orthogonality between the OFDM subcarriers and the
Figure 3.2: Two cycles of MCC time-domain waveform with a 1\(\mu\)s OFDM symbol duration and a 2\(\mu\)s PWM-like envelope in (A) and one cycle showing a PWM-like envelope duration of 10\(\mu\)s transmitting 5 OFDM symbols of 1\(\mu\)s duration each in (B). The spectrum for both cases is shown in (C) and (D), respectively, illustrating the non-overlapping spectrum of OFDM subcarriers, i.e. multi-colored, and 50\% PWM-like harmonics, i.e. red.

PWM-like harmonics is not lost. This is highlighted in Fig. 3.2C where there is no overlap between the PWM-like harmonics and the OFDM subcarriers. An interesting fact is that increasing the number of OFDM symbols per PWM-like cycle can still maintain orthogonality. This is shown in Fig. 3.2B where the positive samples of 5 OFDM symbols of duration 1\(\mu\)s each are shaping the ON period of a PWM-like cycle of duration 10\(\mu\)s while the negative samples are forming the OFF
period. As the PWM-like cycle duration increases from 2\text{us} to 10\text{us}, the fundamental frequency is reduced and so are the harmonics, as shown in Fig. 3.2D. Note that the figures are adjusted, \textit{i.e.} zoomed in, to precisely indicate the relation between the frequency components while the total number of subcarriers is 16.

Unlike the previous cases, Fig. 3.1A illustrates the impact of the number of OFDM symbols per $T_{PWM}$, \textit{i.e.} $C_2$, on the orthogonality loss between PWM-like harmonics and OFDM subcarriers. This is represented by the overlapping frequencies in Fig. 3.1A, where $C_2 = 3$. This scenario arises from improper design considerations resulting in an overlap between PWM-like harmonics and some OFDM subcarriers, such as the odd subcarriers. However, there is no impact of varying the duty cycle, \textit{i.e.} $D$, while maintaining an integer value for $C_1$, on losing the orthogonality. As seen in Fig. 3.1B, the spectral profile shows no overlap between OFDM subcarriers and PWM-like harmonics for three different cases of varying $D$ while $C_2 = 10$. This is reflected on the energy saving, where the communication quality is still maintained while $D$ can be minimal. Both figures are properly zoomed in the same way as in Figs. 3.2C and 3.2D.

### 3.3 Transmitter and Receiver Design

As the MCC unified waveform supports multiple simultaneous modulations, it requires a design procedure that avoids possible interference between the signal components. Additionally, MCC is composed to be supporting multiple receiver complexities. In this section, the transmitter and receiver design is explained. As shown in Fig. 3.3, there are two separate streams of bits to communicate with both high and low speed devices. The high speed bits are OFDM modulated, while the low speed bits are distributed between $L_1$- beacon position modulation ($L_1$-BPM) and $L_2$-beacon phase-shift keying ($L_2$-BnPSK) modulation paths. As to comply with dimming control, MCC is designed upon a PWM-like envelope. According to (3.10, 3.11), a sinusoidal beacon $b(t)$ is $L_2$-BnPSK modulated and sampled; \textit{i.e.} $b_s(t)$, at $T_{PWM}$ intervals. Without loss of generality the beacon cycle, $T_b$, is assumed to be equal to the beacon duration, $T_B$, \textit{i.e.} the embedded beacon...
duration consists of one sinusoidal cycle.

\[ b(t) = \sin \left( \frac{2\pi t}{T_b} + \theta_n \right) \]  

(3.10)

where,

\[ \theta_n = \frac{(2n - 1)\pi}{L_2}, \quad n = 1, 2, \ldots, L_2 \quad \text{for } L_2\text{-BnPSK, } L_2 > 2, \]

\[ \theta_n = (1 - n)\pi, \quad n = 0, 1 \quad \text{for } 2\text{-BnPSK} \]

\[ b_s(t) = \sum_{n=0}^{L_2} b(nT_{PWM}) \delta(t - nT_{PWM}) \]  

(3.11)

In (3.12), beacon PWM sampling is explained, where every beacon sample is represented by a

\[ \text{Beacon} \quad \sim \]

\[ \log_2(L_2) \quad \text{BnPSK Modulator} \quad \text{BnPSK Modulator} \]

\[ \log_2(L_1) \quad \text{PWM Modulator} \]

\[ \text{PWM Envelope} \quad \text{Aggregate} \quad \text{Reshape} \quad \text{DAC} \quad \text{Bias-Tee} \]

\[ \text{High Speed Bits} \quad 010110010111 \]

\[ \text{QAM} \quad \text{Hermitian Symmetry} \quad \text{S/P} \quad \text{IFFT} \quad \text{P/S & CP} \quad \text{Zero Clip} \]

\[ \text{Low Speed Bits} \quad 001011 \]

\[ \text{Bit Splitter} \]

\[ \text{Dimming Level} \]

\[ \text{Figure 3.3: MCC transmitter.} \]

specific PWM duty cycle according to (3.13). The \( L_2\text{-BnPSK} \) modulated PWM-sampled beacon is composed of reshaped ACO-OFDM samples to formulate a PWM-like envelope as shown in (3.13). In Fig. 3.3, the high speed bits modulation path to generate ACO-OFDM samples is shown. The bits are modulated using a quadrature amplitude modulator (QAM) then hermitian symmetry is applied before the inverse fast Fourier transform (IFFT) block to ensure a real output.
All negative samples are clipped to generate positive samples only. Due to the half wave symmetry property of ACO-OFDM, clipping the negative samples only adds clipping noise to the even subcarriers, whereas ACO-OFDM has only the odd subcarriers active. MCC can still be generated using different optical OFDM modulation variations.

\[ b_{PWM}(t) = \sum_{n=0}^{L_2} b_{n-PWM}(t - nT_{PWM}) \]  \hspace{1cm} (3.12)

\[ b_{n-PWM} = \begin{cases} 
2L_1 P + \sum_{k=1}^{i} x_{k-NACO-OFDM}(t), & 0 \leq t \leq T_H(n) \\
\sum_{k=i+1}^{\gamma f - i} x_{k-ACO-OFDM}(t), & T_H(n) < t \leq T_{PWM} 
\end{cases} \]  \hspace{1cm} (3.13)

where,

\[ T_H(n) = \frac{b(nT_{PWM}) - b_{s-min}(t)}{b_{s-max}(t) - b_{s-min}(t)} T_{PWM}, \]

\[ \gamma_f = T_{PWM}/T_{OFDM} \] and \( i \) is the number of OFDM symbols per the on duration of the PWM cycle, \( T_H(n) \). It is essential to note that \( x_{k-NACO-OFDM}(t) \) represents negative ACO-OFDM, where positive samples are clipped rather than the negative samples being clipped as in conventional ACO-OFDM, \( x_{k-ACO-OFDM}(t) \). Reshaping \( x_{k-NACO-OFDM}(t) \) within the on duration of the PWM duty cycle, \( T_H(n) \), allows the OFDM analog samples to be within the PWM dynamic range. Due to the half wave symmetry property, a conventional ACO-OFDM receiver can still be used to detect both \( x_{k-ACO-OFDM}(t) \) and \( x_{k-NACO-OFDM}(t) \) within an MCC envelope [25]. To ensure the orthogonality between OFDM subcarriers and PWM harmonics, the minimum on duration within a PWM sample, \( T_H(n)_{\text{min}} \), of duty cycle, \( d(n)_{\text{min}} \), should allocate an integer number of OFDM symbols as follows,

\[ T_H(n)_{\text{min}} = \frac{d(n)_{\text{min}} T_b}{L_2} = i T_{OFDM} \forall i \in \mathbb{Z}^+ \]  \hspace{1cm} (3.14)

As illustrated in Fig. 3.3, the beacon position is modulated within a PWM-like envelope to represent the \( L_1\)-BPM bits. A conventional PPM envelope follows (3.15), where \( p(t) \) is a unity pulse of
a duration of $T_b$ and $c_k \in \{c_0, c_1, \ldots, c_{L_1-1}\}$ is the PPM symbol sequence, i.e. codeword.

$$x_{\text{PPM}}(t) = \sum_{k=0}^{L_1-1} c_k p(t - \frac{kT_r}{L_1})$$

(3.15)

where $T_r$ represents the envelope duration for $L_1$-PPM offering a bit rate of $R_{b,\text{-PPM}}$ and follows,

$$T_r = \frac{\log_2 L_1}{R_{b,\text{-PPM}}} = L_1 T_b = L_1 L_2 T_{\text{PWM}}$$

(3.16)

Based on (3.15), $L_1$-BPM waveform is explained by replacing $p(t)$ in (3.15) by the PWM modulated beacon, $b_{\text{PWM}}$, in (3.12), as shown in (3.17).

$$x_{\text{BPM}}(t) = 2L_1 P \sum_{k=0}^{L_1-1} c_k b_{\text{PWM}}(t - \frac{kT_r}{L_1})$$

(3.17)

Similar to reshaping the OFDM samples to formulate a PWM-sampled beacon in (3.13), the OFDM samples are reshaped to compose the dimming control PWM cycles as follows,

$$x_{\text{PWM}}(t) = \sum_{n=0}^{M} x_{n-\text{PWM}}(t - nT_{\text{PWM}})$$

(3.18)

$$x_{n-\text{PWM}} = \begin{cases} 
2L_1 P + \sum_{k=1}^{m} x_{k-\text{NACO-OFDM}}(t), & 0 \leq t \leq d_f T_{\text{PWM}} \\
\sum_{k=m+1}^{\gamma_f} x_{k-\text{ACO-OFDM}}(t), & d_f T_{\text{PWM}} < t \leq T_{\text{PWM}}
\end{cases}$$

(3.19)

where $M = T_r/T_{\text{PWM}}$ is the number of PWM cycles per an MCC envelope, $d_f$ is MCC duty cycle responsible for dimming control and $m = d_f T_{\text{PWM}}/T_{\text{OFDM}}$ is the number of OFDM symbols per the on duration of a PWM cycle. Finally, an MCC waveform, $x_{\text{MCC}}(t)$ that the DAC generates; to modulate an LED after using a bias-tee for providing enough DC-bias for the LED operation, is shown in (3.20).

$$x_{\text{MCC}}(t) = x_{\text{PWM}}(t)(1 - x_{\text{PPM}}(t)) + x_{\text{BPM}}(t)$$

(3.20)

One of the main novelties about MCC is that it can support both low- and high-end receivers,
simultaneously. The high-end receivers are based on conventional OFDM reception and processing. A high speed PD captures the optical MCC waveform and converts the signal to the electrical domain followed by a trans-impedance amplifier (TIA). After the signal is digitized using an analog-to-digital converter (ADC) and the cyclic prefix is removed, FFT is used to inverse the IFFT process on the transmitter side. A QAM demodulator captures the high speed bits from the FFT complex output, taking into consideration that only odd subcarriers are active for ACO-OFDM, as shown on the upper side of Fig. 3.4. A simpler low speed and low-complexity receiver is designed to capture low speed bits on the lower side of Fig. 3.4. After the signal is captured by a PD and amplified using a TIA, the high frequency harmonics other than the beacon frequency are filtered using a low pass filter. A low speed PD can filter out the high frequencies, as well. The beacon phase and position within the MCC envelope are then demodulated using both BnPSK and BPM demodulators as demonstrated in Figs. 3.5 and 3.6, respectively. A BnPSK demodulator is built upon the conventional PSK demodulator shown in [77]. The sinusoidal waveform is multiplied $L_2$ times with a sinusoidal waveform with the same frequency but with a phase shift of $2\pi/L_2$ at every multiplier stage, followed by $L_2$ integrator and dump blocks. The maximum output is chosen to indicate the correct phase of the input beacon, representing the BnPSK bitstream. According to [3], a PPM demodulator can be implemented using a matched filter that is matched to the beacon shape that samples every $T_b$ sec and compares the output to a threshold to determine the pulse location using hard decision decoding or using a soft decision decoder and maximum likelihood (ML) detection to determine the received bits. In the proposed approach, a simpler energy detector
Figure 3.5: BnPSK receiver.

Figure 3.6: BPM receiver.

is implemented to measure the energy of the input waveform every $T_b$ sec. The maximum energy indicates the beacon location within the MCC envelope, so the BPM bits are detected.

3.4 Complexity Analysis

An essential feature of MCC is its unique transmitter and receiver design. As illustrated in Fig. 3.3, the transmitter has more blocks than conventional ACO-OFDM to transmit more bits for low-end receivers. For the ACO-OFDM waveform generation, the largest computational complexity is associated with the IFFT block. According to [21], a radix-2 IFFT/FFT algorithm of length $N$ requires $N \log_2(N)$ complex additions, $A_c(N)$, and $N/2 \log_2(N - N + 1)$ complex multiplications, $M_c(N)$. A complex multiplication can be implemented by 4 real multiplications, $M_r(N)$ and 2
real additions, \( A_r(N) \) while a complex addition includes 2 real additions. Therefore, an IFFT/FFT algorithm requires real multiplication operations of \( 2N \log_2(N) - 4N + 4 \) and real addition operations of \( 3N \log_2(N) - 2N + 2 \). Thus, the IFFT/FFT computational complexity is dominated by \( \mathcal{O}(\log(N)) \). According to (3.3), \( L_2 + 1 \) additions and \( L_2 \) multiplications are required to calculate \( T_H(n) \) for \( L_2 \) beacon samples, while the arithmetic complexity of obtaining the minimum or maximum sample is \( \mathcal{O}(L_2) \). As for the PWM-sampled BPM envelope generation based on (3.17), \( L_1 \) multiplications and \( L_1 - 1 \) additions are needed. In order to embed the PWM dimming control duty cycles, 2 additions and 1 multiplication are required as in (3.20). Subsequently, the transmitter complexity is dominated by \( \mathcal{O}(\max(L_2, L_1, \log(N))) \).

Similarly, the high speed receiver is dominated by the FFT block arithmetic complexity, \( \mathcal{O}(\log(N)) \). The interesting fact about the MCC high-end receiver is that it is being built upon the same building blocks of a conventional ACO-OFDM receiver due to the half wave symmetry nature of ACO-OFDM, i.e. \( x_{ACO-OFDM}(n) = -x_{ACO-OFDM}(\frac{N}{2} + n - 1) \), as mentioned in the previous section. Thus, the output of an FFT of the NACO-OFDM symbols the same as an FFT of the original symbols [25]. Additionally, the DC-bias only impacts the DC subcarrier which is not actively carrying any data. As for the BPM receiver, its arithmetic complexity is dominated by the energy detector \( L_1 \) multiplication operations and obtaining the maximum energy, requiring an arithmetic complexity of \( \mathcal{O}(L_1) \). Additionally, a BnPSK receiver requires \( L_2 \) multiplications and integration operations, as well as, obtaining the maximum value after \( L_2 \) integrator and dump blocks. Consequently, the high speed receiver arithmetic complexity is \( \mathcal{O}(\log(N)) \), while for the low speed receivers it is \( \mathcal{O}(\max(L_1, L_2)) \).

### 3.5 Summary

In this chapter, the algorithms for MCC waveform generation and detection are proposed. Additionally, a spectral analysis and interference management study is conducted to investigate the impacts of harmonic distortion due to improper design parameters on the overall BER performance.
Additionally, the transmitter and receiver designs are explained, followed by an end-to-end system complexity analysis. In order to demonstrate a low-end receiver, a LPF was applied, and thus validate the extraction of the sinusoidal beacon signal and system capability of offering a parallel data stream based on single-carrier BPM and BnPSK. A conventional ACO-OFDM receiver is shown to be capable of extracting the high speed bits from an MCC waveform without further modifications. It is concluded that the same MCC waveform could be captured by both high- and low-end receivers, where the relevant information could be captured according to the receiver type.
Chapter 4

MCC Evaluation and Discussion

4.1 MCC Simulation Results

A simulation model is realized in MATLAB to demonstrate the dimming range capabilities of the proposed MCC PHY design to offer parallel high speed data stream, low speed data stream and beacon signals. OFDM samples are realized to generate the PWM-like envelope consisting of the beacon and dimming control parts. The ACO-OFDM signal format is considered to demonstrate the proposed approach; however, the design can be generalized to be integrated with all other forms of unipolar OFDM, for instance, DCO-OFDM. FFT length of 128 is used in the simulation model with 1MHz separation between OFDM subcarriers. The beacon and the PWM-like envelope are generated at frequencies of 20kHz and 100kHz, respectively. A LOS additive white gaussian noise (AWGN) channel is considered. In Fig. 4.1 [1] envelopes with and without OFDM samples are compared to illustrate the effect of varying the SNR on the dimming control range, i.e. FBR boundaries. The SNR is determined by the variance of the Gaussian distributed OFDM samples, i.e. setting the average power per OFDM symbol. The obtained dimming control range is in a decreasing response with values of 80% (10% to 90%), 72% (14% to 86%) and 54% (23% to 77%) for increasing SNR values of 12dB, 18dB and 24dB, respectively. Though there is a clear trade-off between the OFDM signal quality and the dimming range, i.e. SNR and dimming range are inversely proportional, a dimming control swing of 54% of the dynamic range is still achievable at a high SNR value of 24dB.
The BER curves for the OFDM based data stream are obtained at different brightness levels; 5%, 50% and 95% FBR, and as a function of the SNR ranging from 12dB to 20dB. These curves are generated for 256kbits frame size to achieve a BER confidence level better than 95% [78]. As shown in Fig. 4.2, BER values below $10^{-3}$ are obtained for SNR values above 18dB; below the FEC limit of $3.8 \times 10^{-3}$, and the dimming level has no effect on the bit-error performance of the system.

In Fig. 4.3 and Fig. 4.4, the low speed parallel data-stream based on BPM and BnPSK, as well
as, the beacon signal stream are demonstrated. In the upper sub-plots of Fig. 4.3, two frames are transmitted with a single cycle beacon signal at different positions within the whole frame period. A LPF is used to pass the range of beacon frequencies and filter out all other frequency components of the frames. The different light sources could be easily identified by the frequency of the time-domain sinusoids obtained after the LPF; see lower sub-plots of Fig. 4.3. Applying a BPM demodulator; similar to the concept of a PPM demodulation, could determine the location of the beacon within the frame and estimate the associated bits mapped on the beacon position. Binary BnPSK demodulated beacons are also illustrated in Fig. 4.4. The upper sub-plots of Fig. 4.4 show two frames transmitted with a single cycle beacon signal at the same position within the whole frame period, however with 180° phase shift. The lower sub-plots in Fig. 4.4 show the beacons after the LPF, where bits encoded in the phase difference of the beacons could be estimated using a conventional PSK demodulator.

The value of spectral analysis of a designed MCC waveform emerges from the proper spectrum and interference management it enables. For instance, an MCC waveform provides multiple services of dimming control, broadband connectivity on top of sensing and positioning applications,
and thus inter-service interference is to be avoided through spectrum management. Moreover, interference analysis highlights the impact of non-linearity in the front-ends, resulting in an overlap between OFDM subcarriers and PWM-like harmonics, on signal quality and BER performance. In addition, the analysis highlights the impact of external sources of interference, i.e. surrounding light fixtures, whether being only used for lighting or incorporating more services.

In order to support the validity of the analytical work, the BER performance of a 64-QAM MCC waveform is simulated under different interference scenarios in Fig. 4.5. The investigated sources of interference are the ones resulting from light fixtures that offer PWM-based lighting, in addition to those that transmit beacon signals needed for sensing and positioning applications. The deviations in hardware may lead to overlap between OFDM subcarriers and PWM-like harmonics. This overlap has different impacts on the BER performance for different interference scenarios. BER for an MCC waveform without overlapping harmonics and without being exposed to external interference, i.e. in red, shows the best performance, as expected. Being exposed to an external beacon signal that interferes with the PWM-like harmonics only, i.e. no overlap between OFDM and PWM-like harmonics, an MCC waveform still maintains the best BER performance. However, being exposed to the same interfering signal while PWM-like harmonics and OFDM subcarriers
overlap, *i.e.* in gray, requires at least 5dB increase in signal-to-interference-plus-noise ratio (SINR) in order to target BER $\geq 10^{-3}$. It is essential to emphasize that the amplitude of the interfering signal is fixed and independent from the main signal amplitude.

Another source of external interference on MCC waveform with overlapping harmonics is the surrounding light fixtures only used for illumination, *i.e.* PWM dimming control. For the blue curve, the PWM frequency of the interfering waveform is 500kHz, which is the same as the PWM-like envelope of the MCC waveform. However, $D$ of the former is 25% while $D$ of the latter is 50%.

The BER performance is comparable to the scenario of not being subject to external interference. However, varying the PWM frequency of the interfering signal results in more degradation of the BER performance, as shown by the magenta curve. From the no interference scenario, a gap of 2-3dB is obvious at BER of $\geq 10^{-3}$. The constellations in Figs. 4.5B and 4.5C represent the received signal in the same interference scenario, *i.e.* a random beacon signal with the same frequency as one of the OFDM subcarriers, at SINR of 26dB, with and without harmonics overlap, consecutively. The interference effect is observed on the real axis of the constellation due to the real nature of the interfering signal.
Figure 4.5: (A) BER performance for 64-QAM MCC waveform under different interference scenarios, whereas (B) and (C), respectively, show received constellations for a received MCC waveform exposed to an interfering beacon signal at SINR of 26 dB, with and without harmonics overlap, i.e. $C_2=3$ and $C_2=10$.

4.2 Technology Division Multiplexing Trade-offs

MCC is designed to serve multiple technologies simultaneously in what could be considered as technology division multiplexing. One of the main challenges of embedding numerous modulation techniques within MCC is the performance trade-off per technology. This section discusses the allocated power per modulation technique and its impact on BER performance and the overall
spectral efficiency. In Fig. 4.6, the impact of embedding an OFDM waveform within an MCC envelope on the BER performance of BPM is illustrated. Additionally, BPM BER with and without OFDM is benchmarked against OOK and PPM as shown in Fig. 4.6A.
FIGURE 4.6: Impact of embedding OFDM within the MCC envelope on BPM BER.

For fairness in BER performance comparison, both optical power, $P$, and data rate, $R_b$, are unified for all PPM, BPM and OOK modulations. Thus, the optical intensity of an OOK pulse becomes $2P$ for a duration of $1/R_b$; however, for $L_1$-PPM it is $L_1P$ for a duration of $T_{PPM}$. As for $L_1$-BPM,
the required optical intensity to maintain optical power of $P$ is $2L_1P$, where the beacon duration $T_b$ is the same as $T_{PPM}$. According to [79], assuming a high SNR where BER is dominated by the two nearest signals, BER of an AWGN channel of a double sided power spectral density (PSD) of $N_0/2$, zero mean and variance of $\sigma^2$ can be approximated as,

$$\text{BER} \approx Q\left(\frac{d_{\text{min}}}{\sqrt{2N_0}}\right) \quad (4.1)$$

where $d_{\text{min}}$ is the minimum Euclidean distance between any pair of valid modulation signals as shown in (4.2),

$$d_{\text{min}} = \min_{i \neq j} \int (x_i(t) - x_j(t))^2 dt \quad (4.2)$$

Accordingly, OOK and $L_1$-PPM BER behaviors are approximately performing as follows,

$$\text{BER}_{\text{OOK}} \approx Q\left(\sqrt{\frac{E_b}{N_0}}\right) \quad (4.3)$$

$$\text{BER}_{\text{PPM}} \approx Q\left(\sqrt{\frac{E_{b-PPM} \log_2(L_1)}{N_0}}\right) \quad (4.4)$$

A PWM modulated beacon has twice the optical intensity to maintain the same average optical power $P$, thus BPM has twice the $E_{b-PPM}/N_0$ ratio. Therefore, BPM BER performance follows (4.5).

$$\text{BER}_{\text{BPM-NO OFDM}} \approx Q\left(\sqrt{\frac{2E_{b-PPM} \log_2(L_1)}{N_0}}\right) \quad (4.5)$$

As shown in Fig. 4.6A, OOK and 2-PPM BER curves match, while $L_1$-BPM requires 3dB less electrical power to achieve the same BER compared to $L_1$-PPM. For $L_1$-PPM and $L_1$-BPM modulation schemes, the error is confined to the symbol in which it occurs and thus, a single-slot error impacts $\log_2(L_1)$ bits. Assuming full synchronization between the transmitter and a hard decision decoding using a threshold detector in an AWGN channel, the orthogonality between symbols increases as the PPM modulation order $L_1$ increases. As a result, higher PPM modulation orders have better error performance at the same $E_b/N_0$. 

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Embedding an ACO-OFDM within an MCC envelope acts as a noise source to the other modulation types including BPM. According to [24], ACO-OFDM is accurately modeled as a Gaussian random process with zero mean and variance of \( \sigma_{\text{OFDM}}^2 \). Accordingly, there is a performance trade-off between the embedded OFDM waveform and other MCC embedded modulation techniques, including BPM. Subsequently, BPM BER in (4.5) is adjusted to include the impact of embedded OFDM as shown in (4.6), where \( \text{SINR}_{\text{OFDM}} \) represents the OFDM signal to interference and noise ratio including ACO-OFDM waveform clipping beyond MCC envelope as explained in [51]. Fig. 4.6 illustrates that BPM BER deteriorates as OFDM SINR increases. As BPM modulation order \( L_1 \) increases, BPM BER behavior is enhanced for the same OFDM SINR and same BPM \( E_b/N_o \).

\[
\text{BER}_{\text{BPM}} \approx Q\left(\sqrt{\frac{E_b-\text{BPM}/N_o \log_2(L_1)}{1 + 2 \text{SINR}_{\text{OFDM}}}}\right) \tag{4.6}
\]

In [51], a detailed analysis on the optimum ACO-OFDM SINR within an MCC envelope; including the clipping impact, is discussed. \( \lambda_{\text{top}} \), defined as the ratio between the top optical clipping

![Figure 4.7: Impact of MCC-OFDM clipping on required SINR for target BER.](image-url)
level of the ACO-OFDM waveform and its standard deviation, $\sigma_{\text{OFDM}}$, is investigated to decide
the optimum OFDM modulation based on SINR. Using the analysis in [51] to target a FEC BER
threshold of $3.8 \times 10^{-3}$ as a metric of maintaining minimum performance per embedded mixed
carrier modulations, Fig. 4.7 illustrates the relation between $\lambda_{\text{top}}$ and OFDM SINR. As shown in
the figure, increasing the OFDM M-QAM modulation order, requires more SINR for the same $\lambda_{\text{top}}$
to target the FEC BER threshold. The required SINR starts to flatten beyond $\lambda_{\text{top}}$ equals 1.5.
While MCC provides different modulation techniques to serve simultaneous technologies, it is es-
sential to investigate the spectral efficiency gain, $\eta$, MCC offers. According to [80, 81], $L_1$-PPM
bandwidth $\text{BW}_{\text{PPM}}$ can be approximated as follows,

$$\text{BW}_{\text{PPM}} \approx 1/T_b \approx \frac{R_{b-\text{PPM}} L_1}{\log_2(L_1)} \quad (4.7)$$

Thus, PPM is attractive for a range of applications where bandwidth is not of major concern as
power efficiency. However, BPM bandwidth requirements are even more where BPM spectral
efficiency $\eta_{\text{BPM}}$ is constrained by the minimum PWM pulse duration $T_H(n)_{\text{min}}$ sampling the beacon
as shown in (4.8).

$$\eta_{\text{BPM}} = R_{b-\text{BPM}} T_H(n)_{\text{min}} \quad (4.8)$$

In order to investigate the spectral efficiency gain by the low speed modulations embedded within
MCC, i.e. MCC-LS, $L_2$-BnPSK is considered as in (4.9). Without loss of generality, different types
of VLC compatible OFDM modulation techniques can be incorporated; however, this analysis
considers ACO-OFDM. Thus, MCC overall spectral efficiency $\eta_{\text{MCC-(ACO-OFDM)}}$ considering
M-QAM for ACO-OFDM modulation is explained by (4.10), assuming the number of cyclic
prefix subcarriers to be 1/4 of the total subcarriers.

$$\eta_{\text{MCC-LS}} = \frac{\log_2(L_1 L_2) T_{\text{OFDM}}}{L_1 L_2 T_{\text{PWM}}} \quad (4.9)$$

$$\eta_{\text{MCC-(ACO-OFDM)}} = \frac{\log_2(M)}{5} + \frac{\log_2(L_1 L_2) T_{\text{OFDM}}}{L_1 L_2 T_{\text{PWM}}} \quad (4.10)$$
From (4.9) and (4.10), the MCC spectral efficiency gain over conventional ACO-OFDM can be interpreted by (4.11),

$$\eta_g = \log_2(L) \frac{\gamma_f}{L}$$

(4.11)

where $L = L_1L_2$ and $\gamma_f$ is the ratio between $T_{PWM}$ and $T_{OFDM}$. The spectral efficiency gain is pictured in Fig. 4.8. The figure shows the inverse proportionality between $\gamma_f$ and the spectral efficiency gain $\eta_g$. As the BPM and BnPSK modulation orders $L_1$ and $L_2$ increase, $\eta_g$ decreases due to the high bandwidth requirements resulting from reducing the minimum PWM pulse duration $T_{H(n)}_{\text{min}}$. In the following section, the optimum spectral efficiency is analyzed to maximize the MCC spectral gain. As noticed in (4.5), the relation between $\text{SINR}_{OFDM}$ and $\text{BER}_{BPM}$ is inversely proportional. In Fig. 4.9A, the required BPM $E_b/N_0$ to target a FEC BER threshold of $3.8 \times 10^{-3}$ as a metric of maintaining minimum performance for BPM is illustrated with respect to $\text{SINR}_{OFDM}$. It is obvious that increasing the BPM modulation order reduces the power requirements for the same BER target. This is justified by the BPM high power efficiency at the cost of lower spectral efficiency. The behavior in Fig. 4.9A is re-plotted in Fig. 4.9B using (4.12) to illustrate the power
Figure 4.9: Power requirements by MCC for embedding BPM with respect to SINR\textsubscript{OFDM}.
requirements by an MCC envelope to target the FEC BER threshold for BPM.

\[
\text{SNR}_{\text{MCC−LS}} = \frac{E_{0−\text{BPM}} \log_2(L_1)}{N_0 L_1}
\]  

(4.12)

### 4.3 MCC Design Parameters Optimization

**FIGURE 4.10**: Impact of varying PWM frequency on maximum MCC spectral efficiency at target BER, varying MCC-LS and OFDM QAM modulation orders.
As shown in Fig. 4.18, $f_{\text{PWM}}$ is inversely proportional with MCC spectral efficiency gain $\eta_g$. This
indicates as $\gamma_f$ increases, i.e. $f_{PWM}$ decreases, $\eta_g$ is fairly noticeable as BPM and BnPSK modulation orders increase. In order to maximize the spectral efficiency, an optimization problem is formulated and constrained by a BER per modulation of $3.8 \times 10^{-3}$ as follows,

$$
\begin{align*}
\max_{M, L_1, L_2, T_{PWM}} & \quad \eta_{MCC-(ACO-OFDM)} = \frac{\log_2(M)}{5} + \frac{\log_2(L_1 L_2)}{L_1 L_2 T_{PWM} T_{OFDM}} \\
\text{s.t.} & \quad T_r = 150 \text{ms}, \\
& \quad 200 \text{Hz} \leq f_{PWM} \leq \frac{\Delta f}{2}, \\
& \quad \text{BW}_{LED} \leq 10 \text{MHz}, \\
& \quad T_b \geq 2 T_{PWM}, \\
& \quad \text{BER} \leq 3.8 \times 10^{-3}
\end{align*}
$$

The optimization problem is constrained by a number of constraints related to MCC design and performance. This optimization problem is defined as a non-linear problem that is solved using optimization toolbox by MATLAB, based on initialized values.

MCC is designed to support localization on top of communications based on the signal strength of received beacons. Thus, the envelope duration $T_r$ is constrained by 150ms to provide a reasonable advertising interval for reliable positioning accuracy for a walking person within an indoor space. This setting allows a maximum of 300ms advertising time if one beacon is located at the beginning of an envelope while the consecutive beacon is located at the end of the following envelope. According to [82, 83], 100-350ms is an acceptable range for a stable indoor positioning system of a mobile object at a normal walking speed. According to [51], $f_{PWM}$ is constrained by a lower limit of 200Hz to provide flicker-free illumination and an upper limit of $\Delta f$, i.e. $1/T_{OFDM}$, to mitigate PWM harmonics interference with OFDM subcarriers. Also, most commercial LEDs have limited bandwidth not exceeding 10MHz. Additionally, to comply with minimum Nyquist sampling requirements, $T_b$ has to be larger than twice $T_{PWM}$. Subsequently, Fig. 4.10 shows the maximum spectral efficiency obtained by MCC compared to ACO-OFDM at different PWM frequencies, targeting the FEC BER threshold for all modulation techniques. At a lower PWM frequency range
of 200Hz-1kHz, MCC does not show an obvious enhancement in terms of achievable maximum spectral efficiency compared to ACO-OFDM. As \( f_{PWM} \) increases, MCC shows a better performance which is more obvious at the upper PWM frequency limit of 78.125kHz, where MCC with \( L=4 \) offers the highest maximum spectral efficiency compared to higher BPM and BnPSK modulation orders.

In Fig. 4.11, the relation between the required OFDM SINR and MCC SNR to target the FEC BER threshold for BPM modulation; at different BnPSK modulation orders, is illustrated. As shown in the figure, it is clear that the required SNR\(_{MCC−LS} \) increases with the increase of SINR\(_{OFDM} \). As the \( L_2 \) BnPSK modulation order is higher, more SNR\(_{MCC−LS} \) is needed at the same SINR\(_{OFDM} \). However, SNR\(_{MCC−LS} \) requirement; with a range of interest below 40dB, is decreased as \( L_1 \) BPM modulation order is higher. This is highlighted in Fig. 4.12, which shows MCC SNR requirements against OFDM SINR for 8-BnPSK at varying BPM modulation orders to target the BER threshold for 8-BnPSK. In (4.14), BnPSK BER is derived based on (4.1) and MCC structure embedding \( L_1 − BPM, L_2 − BnPSK \) and M-QAM OFDM.

\[
\text{BER}_{BnPSK} \approx Q\left(\sin \frac{\pi}{L_2} \sqrt{\frac{2 \text{SNR}_{MCC−LS} \frac{L_1}{L_2}}{1 + 2 \text{SINR}_{OFDM}}} \right) \tag{4.14}
\]

Hence, it is concluded that increasing both \( L_1 \)-BPM and \( L_2 \)-BnPSK modulation orders results in better power efficiency at the cost of less spectral efficiency gain. In order to maximize spectral efficiency of MCC, the objective function in (4.13) is defined as a non-linear optimization problem. The MATLAB optimization toolbox solver-based approach is utilized to obtain the optimum parameters for maximizing MCC spectral efficiency. Initial values for the problem variables fed into the model are obtained from the plots in Figs. 4.10 to 4.12. Additionally, the BER threshold of \( 3.8 \times 10^{-3} \) is set to be the minimum acceptable BER per MCC technology for solving this problem.

A visualization of the optimum maximized MCC spectral efficiency power requirements is shown in Fig. 4.13. The relation between SINR\(_{OFDM} \) and SNR\(_{MCC−LS} \) is almost linear. As SNR\(_{MCC−LS} \) increases from 4dB to reach 14dB, MCC maximized spectral efficiency range is 2.3-3 bits/sec/Hz. The required SINR\(_{OFDM} \) range is 30-38.5dB. The optimum variables are shown in Fig. 4.14 for the
Figure 4.12: SNR requirements for MCC-LS with respect to SINR_{OFDM} for varying $L_1 -$ BPM modulation orders at target BER for 8-BnPSK.

Figure 4.13: MCC power requirements for spectral efficiency optimization.

The same range of $\text{SNR}_{\text{MCC-LS}}$, i.e. 4-14dB, as well. At this range, the maximum spectral efficiency is achieved by embedding 12-13 bits per BPM symbol, and 2-BnPSK within the MCC PWM-like
envelope of frequency around 78.125kHz, as shown in Fig. 4.14. Using these values, M-QAM modulation order controls the maximum obtained MCC spectral efficiency with a range of 12-15 bits per symbol.

**Figure 4.14:** MCC optimized variables for spectral efficiency optimization.

### 4.4 Summary

MCC is evaluated as a novel VLC PHY transmission that serves receivers with different complexities, simultaneously. Embedding different modulation schemes for technology division multiplexing in a PWM-like envelope comes with energy and spectral efficiency trade-offs. MCC is modeled and evaluated using MATLAB simulation. Simulation results confirm that the BER for the multi-carrier OFDM stream is independent on the dimming level within the useful range. A desired BER performance below $10^{-3}$ for SNR values above 18dB was presented. There is a
trade-off between the SNR of the OFDM stream and the useful dimming range. However, a 54\% dimming range was obtained at 24dB SNR. Additionally, simulation results show an interference-immune performance of a properly managed spectrum with an SNR penalty of 2-5dB for different interference scenarios. Finally, individual formulas for BER per modulation scheme are deduced to set a FEC BER of $3.8 \times 10^{-3}$ threshold as a performance metric to study the design trade-offs. An optimization problem is formulated to maximize MCC spectral efficiency and define MCC optimum variables including modulation orders, PWM frequency and SNR per modulation scheme. A spectral efficiency of around 3 bits/sec/Hz is obtained at MCC SNR of almost 12dB.
Chapter 5

Experimental Evaluation and Preliminary
Results

In this chapter, technology demonstrators and hardware implementations are discussed to investigate the validity of the proposed concepts. The first setup is implementing VLC DCO-OFDM based on commercial-off-the-shelf components. Secondly, MCC is implemented using laboratory optical testbench equipment. More experimental setups related to WiLP is still in early stage of research and to be explained in next chapter.

5.1 DCO-OFDM Adaptive Bit Loading

The majority of exiting VLC communication systems employ commercial-off-the-shelf (COTS) PIN or Avalanche PDs (APD), both offering high SNR and large bandwidth compared to simple PN-type PDs. However, as the VLC technology evolves, lower costs and smaller form factor modules are required to increase the market value. The use of PN PDs in a standard microfabrication technology such as Complementary Metal Oxide Semiconductor (CMOS) puts VLC on the path towards miniaturization and high volume production. However, it also sets bandwidth and SNR constraints that have not been tackled in the majority of published works to date. An experimental investigation of adaptive bit-loading for DCO-OFDM is presented. The optical testing setup is presented in Fig. 5.1. An OFDM signal is generated using MATLAB and fed into the arbitrary
Figure 5.1: VLC DCO-OFDM setup realized over 4m.

wave form generator, AWG 81150A, from Keysight. The OFDM configuration settings are based on channel characterization in the frequency domain to study noise and harmonic distortion and estimate SINR per subcarrier. The total number of FFT points is 32, where 15 of them carry real data and one carries no power. The remaining 16 subcarriers are the complex conjugates of the first 16 ones in order to maintain the Hermitian symmetry required to obtain real signals in DCO-OFDM. The sampling frequency is 250 MHz. Moreover, the signal under experimental test consisted of 3080 bits. To overcome high PAPR and get the full dynamic range of the LED, the signal is clipped at 9.5 dB. The full experimental specifications and OFDM configuration settings are summarized in Table 5.1.

The DCO-OFDM signal is applied to the LED driver circuit. The LED light is steered towards the receiver using a Thorlabs aspheric lens ACL5040U-A (5cm diameter and 4cm focal length). The transmitted power is measured at around 500µW on the receiver side. Another aspheric lens ACL7560U-A (7.5cm diameter, 6cm focal length) is used to focus the light beam over the PD of area 1mm × 1mm. The received signal is measured using the digital oscilloscope DSO9254A from
Table 5.1: DCO-OFDM Setup Specifications and OFDM settings.

<table>
<thead>
<tr>
<th>Experimental setup specifications</th>
<th>Description</th>
</tr>
</thead>
<tbody>
<tr>
<td>Arbitrary waveform generator (AWG)</td>
<td>Resolution: 14 bits</td>
</tr>
<tr>
<td></td>
<td>Sampling rate: 2GSa/s</td>
</tr>
<tr>
<td>Oscilloscope</td>
<td>Bandwidth: 2.5GHz</td>
</tr>
<tr>
<td></td>
<td>Memory depth: 20Mpts</td>
</tr>
<tr>
<td></td>
<td>Sampling Rate: 20GS/s</td>
</tr>
<tr>
<td>LED</td>
<td>Peak wavelength: 650nm</td>
</tr>
<tr>
<td></td>
<td>Cut off frequency: 70MHz</td>
</tr>
<tr>
<td></td>
<td>Capacitance: 6.5pF</td>
</tr>
<tr>
<td></td>
<td>Rise/fall time: 3ns</td>
</tr>
<tr>
<td></td>
<td>Bandwidth: 20MHz</td>
</tr>
<tr>
<td>Photodiode</td>
<td>Responsivity in visible range: 0.16-0.37A/W</td>
</tr>
<tr>
<td>Collimator lens</td>
<td>Diameter: 5cm</td>
</tr>
<tr>
<td></td>
<td>Focal length: 4cm</td>
</tr>
<tr>
<td>Receiver lens</td>
<td>Diameter: 7.5cm</td>
</tr>
<tr>
<td></td>
<td>Focal length: 6cm</td>
</tr>
</tbody>
</table>

<table>
<thead>
<tr>
<th>Design parameters</th>
<th>Description</th>
</tr>
</thead>
<tbody>
<tr>
<td>IFFT/FFT length</td>
<td>32</td>
</tr>
<tr>
<td>Data subcarriers</td>
<td>15</td>
</tr>
<tr>
<td>Sampling frequency</td>
<td>250MHz</td>
</tr>
<tr>
<td>Packet length</td>
<td>3020 bits</td>
</tr>
<tr>
<td>Target BER</td>
<td>$10^{-4}$</td>
</tr>
<tr>
<td>Clipping ratio</td>
<td>9.5dB</td>
</tr>
</tbody>
</table>

Keysight followed by offline processing with Matlab for BER calculation. In order to enhance the performance, adaptive modulation is applied based on bit-allocation algorithm. This algorithm employs the least complex technique of rounding. Depending on the channel frequency response, this algorithm allows individual management of each subcarrier depending on its SINR. In other words, a target BER is set while ensuring it is less than the VLC FEC value of $3.8 \times 10^{-3}$. A modulation order is then chosen for each subcarrier based on the target BER formula for M-QAM modulation. The subcarriers with identical powers have similar modulation orders.

The experimental results discuss the system performance when applying adaptive modulation. In Fig. 5.2, the bit allocation per subcarrier is given based on the channel characterization to study the SINR per subcarrier. Multiple design iterations were conducted to reach a system with high bit rate and acceptable BER. In the shown design, 13 subcarriers are active and no data is transmitted on the first subcarrier, i.e. DC, and the last two subcarriers which exhibit low SINR and thus, high BER. The four boundary subcarriers are allocated with one bit only based on binary-phase-shift-keying (BPSK), while the subcarriers in between, each is allocated with 2 bits, i.e. 4-QAM. In Fig.
it is clear that the estimated SINR per subcarrier puts the usable spectrum, \( i.e. \) beyond 3dB, at around 102 MHz. The spectrum from 23.5 MHz to 85 MHz gives around 60 MHz of usable bandwidth obtained through equalization, \( i.e. \) predistortion. Applying adaptive modulation, enabled an additional bandwidth of 14 MHz on each side of this spectrum using BPSK modulation. In Fig. 5.4, the BER is calculated per subcarrier after sending 3080 bits and comparing to the target BER, \( i.e. \ 10^{-4} \), to investigate the quality of the proposed modulation. From the shown figure, the BER per subcarrier indicates an error free transmission for 11 subcarriers while the subcarriers 2 and 14
Figure 5.4: Received BER per subcarrier.

Figure 5.5: Received constellations per subcarrier.
show a BER of nearly $10^{-2}$. The overall BER is $1.9 \times 10^{-3}$ at a bit rate of approximately 172Mb/s. The received signal constellation per subcarrier is given in Fig. 5.5 which validates the signal quality obtained and BER calculations. In Fig. 5.6, the spectrum of transmitted and received OFDM signals is shown. It is noticed that after 100 MHz, both signals attenuate which is expected. As the frequency increases, the received signal suffers from more attenuation due to the equalization and filtering process. At the DC-level, no power is transmitted or received. The total link DC power consumption is about $345mW$.

### 5.2 Experimental MCC Design Validation

Fig. 5.7 shows the waveform generation in MATLAB. First, beacon signals are PWM modulated based on the IDs of the LEDs. Then, high and low capacity bit streams are separated into two different paths to generate OFDM waveforms and modulate the phases of the beacons. A PWM frame is generated with a fixed duty cycle based on the target illumination level. The beacon is concatenated within the dimming frame based on the BPM low capacity bits. The OFDM samples are then structured to generate a PWM-like envelope, resembling the required PWM frame. The signal is fed into a DAC which modulates a DC-biased LED around a DC-operating point. The
transmitted universal waveform then can be captured by either a high-end receiver or a receiver with limited resources.

Fig. 5.8 shows a proposed receiver design for both the low and high capacity bit streams. The same waveform is captured by both receivers. The high capacity link has a high speed PD while the low capacity link can either use the same PD followed by a LPF circuit or use a low speed PD that can do the filtering process. The filtering process is essential to capture the beacon signal frequency while filtering out all other PWM and OFDM harmonics. Each captured signal goes through a TIA and an ADC for further processing using MATLAB. In software, the two paths are used to capture OFDM bits, IDs of the LEDs and low capacity bits.

An experimental implementation of 64-QAM ACO-OFDM based MCC showing the MCC

<table>
<thead>
<tr>
<th>Table 5.2: MCC experimental specifications and waveform design parameters</th>
</tr>
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<tbody>
<tr>
<td><strong>Experimental setup specifications</strong></td>
</tr>
<tr>
<td>AWG</td>
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<tr>
<td></td>
</tr>
<tr>
<td>Light source</td>
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<tr>
<td></td>
</tr>
<tr>
<td></td>
</tr>
<tr>
<td>high speed receiver</td>
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<td></td>
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<tr>
<td>low speed channel (DC)</td>
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<tr>
<td>Scope</td>
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</tr>
<tr>
<td>Power meter</td>
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<tr>
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<tr>
<td><strong>Design parameters</strong></td>
</tr>
<tr>
<td>$T_f$</td>
</tr>
<tr>
<td>$T_B$</td>
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<tr>
<td>$f_{B(1,2)}$</td>
</tr>
<tr>
<td>$f_s$</td>
</tr>
<tr>
<td>$N_{FFT}$</td>
</tr>
<tr>
<td>$f_{PWM}$</td>
</tr>
</tbody>
</table>
waveform generation and detection is presented in Fig. 5.9. As shown in Fig. 5.9, the OFDM samples are generated and conditioned in a PWM-like envelope in software using MATLAB. The digital samples are then fed into a DAC and biased using a bias-tee, before modulating the light intensity of an LED. The receiver in Fig. 5.9 reverses the process by capturing the light intensity variations using a PD and an amplifier circuit. These variations are demodulated in MATLAB after being digitized using an ADC. The PWM-like envelope is removed and the OFDM symbols are reconstructed, so that the received bits can be extracted. A high power Thorlabs red LED, LED635L, transmits the VLC MCC waveform that is fed through an AWG. A telescopic lens, ACL5049U-A, collects the light beam at 50cm onto a DC-biased PIN PD, PDA10A. The received signal is captured using an oscilloscope and post processed using MATLAB where the signal is reconstructed and the received signal and noise samples are averaged to measure SNR. Then, BER is calculated from the relation between SNR, error vector magnitude (EVM) and BER for $M$-QAM published in [84]. The SNR is varied by varying the transmitter power in addition to the distance between the transmitter and the receiver. Fig. 5.9B shows the received signal before being post processed to capture the OFDM modulated bits.

The setup in Fig. 5.10 illustrates the high and low capacity links based on the receiver used to capture the transmitted waveform. The received waveform is captured by an amplified detector, PDA10A, to capture the OFDM bits. The DC-out channel of a receiver module, New Focus 1601, is used to capture the low capacity bits embedded within the transmitted waveform. The received waveforms from both branches are captured using an oscilloscope and then further processed using MATLAB. The experimental setup and waveform design specifications are summarized in Table 6.1.

The trade-off between the OFDM signal quality and dimming capability is captured in Fig. 5.11. Here, the SNR of the generated OFDM samples are varied between 12dB, 18dB and 24dB. A full PWM frame is generated with an embedded PWM encoded beacon and a fixed duty cycle of 50%. The x-axis represents the received signal power, $P_{rx}$, of a PWM frame without OFDM samples while the y-axis represents the signal power of a PWM-like envelope generated by OFDM
The received power is measured at a fixed distance of 50cm using Thorlabs PM100USB power meter and S121C PD sensor. Fig. 5.11A shows a slight variation between the received power of the PWM frame without OFDM and the PWM-like frame with OFDM at SNR of 12dB. This is highlighted in Fig. 5.11B which shows the reduction in dimming range control due to the reduction of PWM-like peak-to-peak voltage as a result of OFDM sample generation. As the SNR increases, the variation in the received power increases. The obtained dimming control range for SNR of 12dB is nearly 93% where the received signal power of a PWM-like envelope with OFDM, i.e., y-axis, varies from 0.51mW to 2.45mW compared to a range between 0.51mW to 2.6mW for a PWM frame without OFDM. Similarly, dimming control ranges of 87% and 60% are obtained for SNR values of 8dB and 24dB, respectively. Though, there is a clear trade-off between the OFDM signal quality and the dimming range. SNR and dimming range are inversely proportional; dimming control swing of 60% of the dynamic range is still achievable at a high SNR value of 24dB.

Next, the high speed link is evaluated based on calculating the BER from the relationship among SNR, EVM and BER for $M$-QAM published in [84]. The SNR is measured at varying steps over a 50cm distance between the transmitter and receiver. According to the limitation on LED dynamic range, the maximum obtained SNR of OFDM samples conditioned to form the PWM-like frame is 26dB, as shown in Fig. 5.12. As expected, the obtained SNR values result in a reasonable BER performance for different modulation orders. For instance, a BER of $7 \times 10^{-5}$ is obtained for a 64-QAM modulated waveform.

In order to evaluate the low capacity link, the low speed receiver is used to capture the waveform and compare it with the high speed one. In Fig. 5.13 the embedded beaconing functionality of the designed waveform is shown. The same transmitted waveform is captured by the high and low speed receivers as shown in the figure. In Fig. 5.13A the PWM-like encoded beacon, OFDM samples and fixed PWM-like duty cycles for dimming control are illustrated, unlike Fig. 5.13B where all PWM-like and OFDM harmonics are filtered out and only the beacons are captured. In
**Figure 5.7:** Proposed system transmitter illustrating waveform generation by MATLAB, as well as, the hardware blocks.

**Figure 5.8:** Proposed system receivers for both low and high capacity bits embedded within the received waveform.
FIGURE 5.9: (A) Experimental setup of ACO-OFDM based MCC for a 50cm separation between a high power LED and a DC-biased PIN PD, (B) Received signal screenshot showing different design parameters, and (C) Experimental BER of reconstructed and demodulated OFDM samples on MATLAB for a 64-QAM.
FIGURE 5.10: Experimental setup to illustrate the reception process using two different receivers for high and low capacity bit streams in (A) and (B), respectively.
Figure 5.11: Experimental impact of varying SNR of conditioned OFDM samples on the dimming capability of the transmitted waveform frame in (A) and effect on dimming control range in (B).
**Figure 5.12:** Experimental BER of reconstructed and demodulated OFDM samples.

**Figure 5.13:** (A) Captured frame by the high speed receiver illustrating the PWM encoded beacon, OFDM samples and fixed 50% duty cycles for dimming control. (B) Captured 20kHz beacon signals by the low speed receiver after concatenation from different frames.
Figure 5.14: Captured frames with embedded PSK modulated beacons using a high speed receiver in (A) and (B) and a low speed receiver in (C) and (D).
Figure 5.15: Captured frames with embedded BPM modulated beacons using a high speed receiver in (A) and (B) and a low speed receiver in (C) and (D).
this figure a screen-shot of the received waveform by the low speed receiver, with 20kHz embedded beacon, is shown after concatenation of multiple beacon signals. The demodulation capability of low capacity bits using a low speed receiver is clear from Figs. 5.14 and 5.15. In Fig. 5.14 an example of a BPSK modulated waveform is illustrated. To the left a frame with 0° phase shift PWM encoded beacon, i.e. “0” bit, is captured by a high speed receiver while to the right a 180° phase shift is received. In Figs. 5.14C and 5.14D the low speed receiver easily captures the beacon’s phases. A BPSK demodulator can be easily applied in MATLAB to capture the bit stream. Similarly, BPM modulated beacons are captured within their designated time slots in Fig. 5.15. On the left, the beacon is placed within the first time slot to represent a “000” bit pattern while the beacon representing “011” is on the right side. In both cases, the low speed receiver successfully filters out the PWM-like and OFDM harmonics without corrupting the beacon’s phase or position within the frame.

5.3 Summary

In this chapter, hardware implementations for DCO-OFDM bit-loading, followed by MCC validation are presented. The limitations in bandwidth at the receiver side due to the use of Si PN PDs are addressed using a custom receiver design with post equalization. Measurement results show a data rate of 172Mb/s at BER of 1.9 × 10^{-3} from a 4m distance VLC link that uses DCO-OFDM and a Si PN PD with an intrinsic bandwidth of 40MHz. Further investigations are being conducted to realize a nearly flat channel with higher SINR which would, in addition to accurate channel estimation and power allocation, lead to higher bit rates.

The experimental investigation of the MCC design framework is presented. An experimental evaluation is performed to demonstrate and validate the simultaneous services offered by MCC; high speed communication, low speed communication, beaconing and dimming control. Based on the experimental results, the proposed technique can offer dimming control over 60% of the LED dynamic range while maintaining a reliable communication link for a 64-QAM at BER of 7 × 10^{-5}.
Chapter 6

Futuristic Concepts: Assistive Links for IoTs and PHY Security

6.1 Introduction

Indoor localization and sensing are interesting VLC deployments to achieve very high level of accuracy with lower cost compared to RF techniques such as WiFi or ultra-wideband (UWB). In such deployments, a unique beaconing signal, \textit{i.e.} mainly a sinusoidal waveform, is transmitted from individual LEDs \cite{85, 86}. The VLC technology integrated with the emerging technology of IoT opens up to wide range of indoor applications such as machine-to-machine (M2M) communication. VLC links could still be maintained with extremely low level of luminance using highly dimmed LEDs, where LEDs are made to appear off \cite{24, 87}. Such deployment consumes much lower energy compared to conventional VLC deployment, making it more suitable for IoT applications. In addition, other VLC deployments are designed with multiple functionality to support high speed communications, as well as localization and sensing applications \cite{72}. Many countermeasures are proposed including the use of an array of LEDs to construct the OFDM signal by summing transmitted signals from the different LEDs of the array in the optical domain rather than generating the signal in the electrical domain. H. Dong et al. propose an optical-domain generated DCO-OFDM signal \cite{88}. Their approach aims at reducing signal clipping distortion in Optical OFDM systems by using multiple LEDs. Each LED transmits
a sinusoidal signal to represent an individual subcarrier and the OFDM signal is formed from the superposition of the optical power from the array of LEDs. However, the received signal remains the same compared with conventional DCO-OFDM systems. More studies on spatial summation of optical power from multiple LEDs are introduced [32, 89]. In their analysis, the authors introduce spatial optical OFDM which divides a wideband input signal amongst multiple LED strings in a luminary. Although this technique is claimed to mitigate the optical OFDM PAPR problem, the receiver still has to be a conventional optical OFDM receiver. It is worth mentioning that the core operation of an OFDM modulator and demodulator are the IFFT and FFT, respectively. Such operations are complex for resource limited IoT devices and complex relative to the operations used in single-carrier pulsed schemes such as OOK.

Moreover, According to [48], hybrid wireless networks are foreseen to play a major role in the visioning and planning of the 6G network. Hence, security and privacy should be key factors in the design of the 6G wireless network. Eavesdropping attacks, a prerequisite of other malicious attacks, have become one of the major threats. Currently, cryptographic algorithms are being deployed for enhanced systems security. However, there are difficulties and vulnerabilities in secret key distribution and management. Recently, PHY security has become an emerging area of research for enhanced security.

Among the highlighted strengths of optical wireless communications (OWC), security is usually listed at the top [7]. Leveraging the fact that light emitted by LEDs is directional, VLC or LiFi networks are envisioned to be more secure than RF networks. For example, the level of security of the network can be increased due to the physical nature of light waves and the ease to confine signals within a single room [90]. However, eavesdropping security concerns still exist and few studies have been done regarding the intrinsic security of VLC, specifically in public areas.

Against indoor eavesdropping attacks, the use of null-steering, artificial noise (AN) and friendly jamming strategies are usually considered. In [91], a portion of the optical signal is dedicated to the message while the other portion is transmitting known jamming signals to degrade the
eavesdropper’s SNR, without affecting legitimate users. Similarly, LEDs surrounding the transmitting LED source are used in [92] to act as jammers that can affect the eavesdropper’s received signal quality. Possibility of outdoor eavesdropping attacks through keyholes and door gaps are also considered [93]. A limited number of publications tried to determine secrecy boundaries for indoor optical wireless systems [94, 95].

Recently, in the majority of hybrid networks, one medium is used at a time. Accordingly, most of the integration efforts are directed towards opportunistic offloading or switching of traffic [8]. For example, hybrid wireless local area network (WLAN) systems coexisting LiFi with legacy WiFi are still vulnerable to channel dynamics, different mobility patterns, as well as, security attacks. The performance of individual links and the security requirements are still managed separately. To the best of the authors’ knowledge, a joint RF-optical security approach does not exist.

Nowadays, most mobile devices are equipped with multiple wireless interfaces. The cognitive radio technology also supports fully programmable and dynamically adaptive transmission waveforms, channel access methods, and networking protocols [96]. Additionally, parallel processing is employed and multiple antennas are necessary for multiple-input and multiple-output (MIMO) and multi-bands operations [97].

### 6.2 Novelty and Contribution

In this chapter, two futuristic novel concepts based on assistive links are proposed.

- **LwO-OFDM** is proposed to support low-end receivers for IoTs using cooperative access points. An array of light sources to realize LwO-OFDM; as a compatible version of the communication chain with resource limited IoT devices, is studied. In the proposed concept, different sinusoidal streams from coordinated LED light sources are carrying the encoded OFDM time-domain samples, thus enabling the realization of FFT in the optical domain.
Accordingly, the FFT operation required for the demodulation at the receiver side is eliminated. Such FFT-less demodulation enables resource limited IoT devices to receive OFDM transmission. In addition, the proposed concept, (1) offers the same spectral efficiency as the well-known ACO-OFDM, (2) reduces the bandwidth requirement from individual light sources, (3) reduces the PAPR of the signal formed and transmitted over the optical channel, and (4) supports simultaneous sensing applications using the different sinusoidal streams that are acting as unique beacon signals. Furthermore, the proposed concept supports reduced complexity channel estimation and equalization for IoT devices.

- **WiLP** is motivated by the envisioned device capabilities and the possible interplay between several access technologies. This concept investigates the potential of waveforms from different air interfaces assisting in securing a hybrid wireless network and filling in the security gap when switching between LiFi and WiFi modes of operation. The envisioned hybrid network ensures outstanding security not through the intrinsic nature of the optical link propagation but the intrinsic characteristic of WiLP. Therefore, WiLP will enhance the wireless network security performance through the dependency on a joint RF-optical design and simultaneous transmission. The reception of one of the links will not be sufficient for an eavesdropper to decode the transmitted information. Additionally, in cases where there will be unauthorized access to both links, the RF link within the multi-air-interface network will act as a AN source or can actively jam back an eavesdropper anticipating more security and higher degree of heterogeneity.

### 6.3 LwO-OFDM

#### 6.3.1 System Overview and Analysis

The structure of a transceiver supporting the proposed LwO-OFDM system is shown in Fig. 6.1. For the transmitter, comparing with the conventional ACO-OFDM system using a single LED as
shown in Fig. 1.2, the building blocks to generate the time-domain samples $X[n]$ are identical, i.e. in blue. After the IFFT block, due to the anti-symmetry of an ACO-OFDM signal, the proposed LwO-OFDM only requires the transmission of the first $\frac{N}{2}$ time domain samples. These samples are used to scale the amplitudes of different sinusoidal signals. The frequencies of these signals are set based on the subcarrier frequencies and subcarrier separation of the system. These sinusoidal signals are used to modulate the intensities of an array of LEDs. To transmit a single OFDM symbol of period $T$, the $\frac{N}{2}$ samples are transmitted over two symbols, where $X_a(t)$ is transmitted at $T_a$ and $X_b(t)$ is transmitted at $T_b$. During $T_a$ and $T_b$, i.e. normally equal, samples are carried over cosine and sine signals, respectively. A phase shift of $180^\circ$ is added to ensure the proper integration of the negative sign before the imaginary component as shown later in the following analysis in (6.3). A bank of $\frac{N}{2} - 1$ oscillators is phase controlled by a $270^\circ$ phase shift controller to generate cosine and negative sine signals with proportional amplitudes to the IFFT time vector $x[n]$. There is no need for an extra oscillator to represent the $0^{th}$ subcarrier which represents the DC-bias of the signal, as it can be replaced by proper DC-biasing of the remaining LEDs. Each sinusoidal waveform modulates one of the LEDs array and then the optical power from the LEDs is summed on air. As the signal is generated with the above considerations, the summation process resembles the FFT process in conventional ACO-OFDM receivers, i.e. FFT-on-air.

In order to have better insight over the transmitted signal shape, the digital samples $x[n]$ after an IFFT of length $N$ on the transmitter side can be represented by (6.1).

$$x[n] = \frac{1}{\sqrt{N}} \sum_{k=0}^{N-1} X[k] \exp\left(\frac{j2\pi k n}{N}\right), \quad \text{for } 0 \leq n \leq N - 1 \quad (6.1)$$

where, the first $\frac{N}{2}$ samples of the vector $X_k$ represent the output samples of the mapper, i.e. QAM-modulator, and the zero-forced even subcarriers as in conventional ACO-OFDM. The following $\frac{N}{2}$ samples are complex conjugates of the first $\frac{N}{2}$ samples in order to ensure real, i.e. non-complex, valued IFFT output samples. Conventionally, an OFDM receiver needs to perform FFT to get IQ samples for a QAM demodulator. The $\hat{X}[k]$ vector is obtained from the FFT process.
FIGURE 6.1: Proposed LwO-OFDM block diagram with common blocks of ACO-OFDM in blue and presenting FFT-on-air process and FFT-less receiver design.
shown in (6.2).

\[
\hat{X}[k] = \frac{1}{\sqrt{N}} \sum_{n=0}^{N-1} x[n] \exp\left(\frac{-j2\pi kn}{N}\right), \quad \text{for } 0 \leq k \leq N - 1
\] (6.2)

Inspired by (6.2), the proposed technique promotes FFT-less receivers to be compatible with IoT and sensing applications. In this context, \( \hat{X}[k] \) can be rewritten as

\[
\hat{X}[k] = \frac{1}{\sqrt{N}} \sum_{n=0}^{N-1} x[n] \left(\cos\left(\frac{2\pi kn}{N}\right) - j\sin\left(\frac{2\pi kn}{N}\right)\right), \quad \text{for } 0 \leq k \leq N - 1
\] (6.3)

As clearly shown by (6.3), using the bank of oscillators in LwO-OFDM enables applying FFT-on-air. This enables the reception of an OFDM signal with a simple FFT-less receiver. The transmitted signals follow what is indicted in Fig. 6.1, where analog signals \( \hat{X}_a(t) \) and \( \hat{X}_b(t) \) of durations \( T_a \) and \( T_b \) are detected by the PD to represent the transmitted real and imaginary coefficients of \( \hat{X}[k] \), i.e. should be detected in a conventional ACO-OFDM system, respectively. These analog signals are expressed in (6.4-6.5).

\[
\hat{X}_a(t) = \frac{1}{\sqrt{N}} \sum_{n=0}^{N-1} x[n] \cos(2\pi f_n t), \quad \text{for } 0 \leq t \leq \frac{T}{2}
\] (6.4)

\[
\hat{X}_b(t) = \frac{1}{\sqrt{N}} \sum_{n=0}^{N-1} x[n] \sin(2\pi f_n t), \quad \text{for } \frac{T}{2} \leq t \leq T
\] (6.5)

where, \( T \) is the LwO-OFDM symbol duration and \( T_a \) and \( T_b \) are both equal to \( \frac{T}{2} \). The received signals are then digitized as in a conventional ACO-OFDM receiver using an ADC to convert \( \hat{X}_a(t) \) and \( \hat{X}_b(t) \) into \( \hat{X}_a[k] \) and \( \hat{X}_b[k] \), respectively. After that, \( \hat{X}[k] \) is reconstructed as in (6.6).

\[
\hat{X}[k] = \hat{X}_a[k] + j \hat{X}_b[k], \quad \text{for } 0 \leq k \leq N - 1
\] (6.6)
FFT-on-air process can be shown as described above in Fig. 6.2. From the previous equations, it is noted that LwO-OFDM is designed to have a symbol duration $T$. However, in order to transmit the real and imaginary coefficients on two different time slots the total symbol time should be double that of a conventional ACO-OFDM. This will reduce the spectral efficiency to half of ACO-OFDM.

An interesting property of the proposed scheme can be illustrated in Figs. 6.2A and 6.2B. In Fig. 6.2A the obtained digital signal, *i.e.* $\hat{X}_a[k]$, from the spatial summation of the sinusoidal, *i.e.* cosine, waveforms is symmetric. In other words, the second $\frac{N}{2}$ samples are the image of the first $\frac{N}{2}$. This is also the case in Fig. 6.2B but with the image inverted. In fact, it is waste of resources to send these redundant images, so our technique is adjusted to send only half of the time samples as in Figs. 6.2C and 6.2D. As a result the whole symbol duration of LwO-OFDM is the same as a conventional ACO-OFDM. As previously discussed, an ACO-OFDM signal has an anti-symmetry shape. In a conventional ACO-OFDM system, the redundant negative values are clipped in order to obtain higher power efficiency. This is illustrated in Fig. 6.3A in blue.
In LwO-OFDM, the oscillators are phase controlled which enables the generation of sinusoidal waveforms representing the negative values from the IFFT, as clear in Fig. 6.3 in red. Therefore the zero-clipping process can be eliminated and replaced by sending half of the subcarrier samples only, i.e. $\frac{N}{2}$, to avoid redundancy. This has a direct benefit on the consumed bandwidth by each single LED, which becomes half of the needed spectrum by an LED transmitting a conventional
ACO-OFDM signal that carries the same data. This is clearly presented in Fig. 6.3B with a simulated LwO-OFDM consuming 4MHz while ACO-OFDM that carries the same data consumes 8MHz. The simulation parameters to generate this curve are detailed in the following section. From the above it is clear that LwO-OFDM transmits one symbol over two time slots, \( T_a \) and \( T_b \), each with half the number of IFFT coefficients \( i.e. \frac{N}{2} \). Interestingly, this results in obtaining a symbol duration equal to that of an ACO-OFDM and same number of active and total subcarriers, as well, \( \frac{N}{4} \) out of \( N \). As expected, this results in the same spectral efficiency of a conventional ACO-OFDM as shown in (6.7). Although, the CP is not discussed in this manuscript, it is still valid to apply sacrificing more oscillators.

\[
R_{LwO} = \left( \frac{\frac{N}{4}}{N + N_{cyc}} \right) \log_2 M \quad \text{bits/s/Hz} \quad (6.7)
\]

### 6.3.2 Simulation Results

A simulation model is realized in MATLAB to demonstrate the performance of the proposed scheme compared to the conventional ACO-OFDM systems. However, the proposed scheme can be adopted using other forms of optical OFDM techniques. An IFFT of length \( N=16 \) is exploited in the simulation model with a subcarrier separation of 1MHz. The sampling frequency \( f_s \) complies with the Nyquist rate \( i.e. \) double the maximum frequency) with a value of 16MHz. On account of the fact that the complex conjugate amplitudes after the IFFT are not transmitted, only 7 light sources and oscillators \( i.e. \frac{N}{2} - 1 \) are needed on the transmitter side. This number can be increased to 8 while the first LED will only be responsible for DC-bias \( i.e. \) 0th subcarrier), but in our case it can be replaced by biasing all the light sources in the LED array with the appropriate DC-bias. BER is simulated for conventional ACO-OFDM and LwO-OFDM using different modulation orders (32-256 QAM) in order to study the effects of SNR and clipping on the transmitter side on the system performance. Higher QAM orders are targeted to compensate for the low spectral efficiency. Table 6.1 summarizes the simulation and system design parameters.
Figure 6.3: (A) Time domain coefficients of zero-clipped conventional ACO-OFDM, \textit{i.e.} in blue, and LwO-OFDM subcarriers, \textit{i.e.} in red, and (B) reduced bandwidth requirements for LwO-OFDM compared to conventional ACO-OFDM due to sending half of the subcarriers \((\frac{N}{2})\).
In Fig. 6.4, there is a clear enhancement in BER performance when LwO-OFDM is compared to conventional ACO-OFDM. In low SNR values till about 23dB, the LwO-OFDM performance is nearly the same as the ACO-OFDM. For a fair comparison, the power of an ACO-OFDM symbol is divided on both real and imaginary spatially summed symbols of LwO-OFDM. As expected, when the QAM order increases the performance degrades and higher SNR levels are needed to achieve the same BER as in lower QAM orders. Both curves of LwO-OFDM and ACO-OFDM approach a BER of nearly $10^{-4}$ at a range of 23dB to 32dB for 32-256 QAM orders. The regions where the performance of both systems is matching are the regions where the there is nearly no clipping or clipping effect on the transmitted signals is negligible. As the SNR exceeds 23dB, signal clipping starts to degrade the performance due to the increased signal power above the LEDs dynamic range. Unlike LwO-OFDM, the clipping effect is severe and quick for ACO-OFDM signals due to relatively higher PAPR. For instance, there is a gain of 10dB-12dB in different modulation orders for the favor of the proposed scheme. There is a good agreement between the obtained results and the signal generation procedure for both techniques. In an ACO-OFDM system, a single LED emits an OFDM symbol with high PAPR characteristics while in LwO-OFDM, multiple

<table>
<thead>
<tr>
<th>BER</th>
<th>SNR (dB)</th>
</tr>
</thead>
<tbody>
<tr>
<td>10^-4</td>
<td>23dB-32dB</td>
</tr>
</tbody>
</table>

**Figure 6.4:** BER of conventional ACO-OFDM, *i.e.* blue, and LwO-OFDM, *i.e.* red, at different SNR levels and showing clipping effects on the signal performance for (A) 32-QAM (B) 64-QAM (C) 128-QAM (D) 256-QAM.
<table>
<thead>
<tr>
<th>System design specifications</th>
<th>ACO-OFDM</th>
</tr>
</thead>
<tbody>
<tr>
<td>OFDM technique</td>
<td>ACO-OFDM</td>
</tr>
<tr>
<td>IFFT length (N)</td>
<td>16</td>
</tr>
<tr>
<td>subcarrier separation</td>
<td>1MHz</td>
</tr>
<tr>
<td>Sampling frequency ($f_s$)</td>
<td>16MHz</td>
</tr>
<tr>
<td>QAM order</td>
<td>32, 64, 128, 256</td>
</tr>
<tr>
<td>Symbol duration ($T$)</td>
<td>1us</td>
</tr>
<tr>
<td>Number of oscillators/light sources</td>
<td>7</td>
</tr>
</tbody>
</table>

<table>
<thead>
<tr>
<th>Simulation parameters</th>
<th></th>
</tr>
</thead>
<tbody>
<tr>
<td>LED dynamic range</td>
<td>0.5V</td>
</tr>
<tr>
<td>Noise level</td>
<td>-10dBm</td>
</tr>
<tr>
<td>SNR range</td>
<td>10:40dB</td>
</tr>
<tr>
<td>Number of symbols</td>
<td>10000</td>
</tr>
</tbody>
</table>

LEDs are responsible to generate sinusoidal waveforms with fixed PAPR of nearly 3dB that sum up on air. These results indicate that LwO-OFDM signal is more resilient to light source clipping. The PAPR comparison between both techniques is given by the CCDF of PAPR which denotes the probability that the emitted signals from LEDs exceed a certain threshold $\text{PAPR}_o$ and is given by (6.8)

$$\text{CCDF} = P(\text{PAPR} > \text{PAPR}_o)$$

where, a general expression for PAPR of a transmitted signal $x[n]$ can be deduced from (6.9)

$$\text{PAPR} = \frac{\max |x[n]|^2}{E[|x[n]|^2]}$$

As previously stated, the LwO-OFDM symbol is mainly formed from the summation of the optical power of sinusoidal waveforms on air. Therefore, in (6.8) $x[n]$ is replaced by $x[n]\cos(2\pi f_s t + \phi)$. As a result, the proposed scheme shows 5dB PAPR reduction, i.e. fixed PAPR of nearly 3dB, compared to ACO-OFDM at CCDF of $10^{-2}$ which is clearly illustrated in Fig. 6.5.
6.4 WiLP

6.4.1 Proposed System

WiLP is considered as an intermediate mode within hybrid networks, which can fill in the security gap during the handover between different modes of network co-existence. For instance, WiLP can prevent out of room intruders from eavesdropping or one of the WiLP paired links could be used to send security keys to the receiver. The joint processing of RF and optical signals can physically add another layer of security to the communication link. Moreover, one link could be used to actively attack a jammer or an eavesdropper’s malicious network intervention. Bits can be categorized and transmitted on either the RF or the optical link relying on the targeted quality of service. Alternatively, OFDM subcarriers can also be mapped on either the RF or the optical link allowing efficient and flexible utilization of the resources depending on whether reliability or high throughput is to be favored. Additionally, each link could have its own adaptive transmission configuration, i.e. modulation order, coding rate, and more importantly transmitted power based on the relative channel conditions. Such flexible adaptation avoids data rate bottlenecks impacted by the poorest channel conditions of the RF-optical links.

In Fig. 6.5, the building blocks of one of the possible realizations of WiLP system are shown. The
system is built upon the IEEE 802.11 standards of PHY specifications and WLAN packet design [98]. The transmitted packets are redesigned to be simultaneously sent over both RF and optical links. The redesign process allows a legitimate user inside a room, who can better receive both signals, to estimate the transmitted data. An eavesdropper located outside the room and physically blocked from receiving the optical signal, can not reconstruct and process the received stand-alone RF signal.

The RF-optical link dependency through WiLP starts from separating the real and the imaginary parts of a QAM modulator. The real components are transmitted over the RF link, while the imaginary components are transmitted over the optical link. The WLAN packet structure is modified to accommodate the proposed changes. For instance, both RF and optical OFDM symbols have binary phase-shift keying (BPSK) based pilots to facilitate phase corrections. In both branches, short and long preambles are added within the transmitted packets for synchronization and channel estimation and equalization. The optical signal modulation is made either in baseband or over a carrier using an IQ modulator. The RF signal is modulated over a WiFi carrier in the GHz band. The legitimate user should be able to receive both RF and optical signals, concurrently. The captured packets from both branches should be handled separately to perform channel estimation, phase correction and equalization due to the different nature of the RF and optical channels. Based on IEEE 802.11 standards, the exponential channel model is adopted to represent a 2.4GHz channel, where the average channel power, \( P(\tau) \), decreases exponentially with the channel delay, \( \tau \), as follows,

\[
P(\tau) = \frac{1}{\tau_d} \exp\left(-\frac{\tau}{\tau_d}\right)
\]

(6.10)

where, \( \tau_d \) represents the root mean square (RMS) delay spread [98]. Therefore, the received RF packet suffers from multi-path fading and requires proper equalization to restore the transmitted bits correctly. On the other hand, the optical channel characteristic, as previously explained in Chapter 1, is considered a quasi-static channel. Thus, the optical channel is dominated by the LOS propagation model with minimal attenuation and scattering.

The design is implemented and validated in MATLAB using the WLAN toolbox. The whole
Figure 6.6: WiLP transmitter and receiver block diagrams. Both RF and optical links use an IQ modulator. The optical link can also be transmitted in baseband.
system is implemented according to the parameters in Table 6.2. As shown in Fig. 6.7, an experimental setup is implemented to test the performance of WiLP. The formulated packets are designed using MATLAB and generated based on the software-defined radio (SDR) technology using universal software radio peripheral (USRP)-B210 boards. The packets are separated on RF and optical links. The distance between the transmitter and the receiver is set at 50cm. The optical packets are sent via LD through a collimator lens and captured by ACL5040U concentrator lens onto the surface of Thorlabs PDA015A PIN PD module. The LD modulation bandwidth is 100MHz and the range of frequency operation of USRP-B210 boards starts at 75MHz. Accordingly, 5MHz packets are generated and the telescopic antennas are used to send the RF packets at a carrier frequency of 85MHz.

The received packets from both RF and optical links are decapsulated to be processed, separately. Both RF and optical channels are estimated based on long preambles and pilots. The equalization process is performed for both RF and optical constellations, on the real and imaginary axes, concurrently. The one-dimensional constellations are mixed up to form the original two-dimensional $M$-ary constellations. The transmitted bits are estimated and the ratio of how many bits received in error over the number of total bits received is calculated to obtain the BER. The BER is obtained for different modulation and coding schemes (MCS) as summarized in [98] (Table 17-4). To measure the security performance aspect of the design, an eavesdropper is assumed to be located outside

<table>
<thead>
<tr>
<th>Simulation Parameters</th>
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</thead>
<tbody>
<tr>
<td><strong>Distance</strong></td>
</tr>
<tr>
<td>RF carrier frequency</td>
</tr>
<tr>
<td>Bandwidth</td>
</tr>
<tr>
<td>Modulation and coding schemes (MCS)</td>
</tr>
<tr>
<td>Packet length</td>
</tr>
<tr>
<td>Number of packets</td>
</tr>
<tr>
<td>Photodetector active area</td>
</tr>
</tbody>
</table>

**Table 6.2: WiLP simulation parameters.**
Figure 6.7: Experimental setup of WiLP at a distance of 50cm between the transmitter and the receiver.

the room and being able to only receive the RF signal. Such reception of partial packets is investigated using the same testbed and the BER performance is determined and compared to that of a legitimate user capturing the packets from both RF and optical links. Perfect communication secrecy is achieved when the legitimate user can communicate reliably and confidentially, while the eavesdropper gets zero bits of information. The link secrecy is measured by the secrecy capacity, \( C_{\text{sec}} \), which is defined as the maximum transmission rate at which the eavesdropper is unable to decode any information. It is equal to the difference between the the main channel and the wiretap channel capacities, as shown in (6.11).

\[
C_{\text{sec}} = C(ch_U) - C(ch_E) = H(P_e(E)) - H(P_e(U)) \tag{6.11}
\]

where, \( C(ch_U) \) and \( C(ch_E) \) are, respectively, the channel capacities of the user’s and eavesdropper’s channels, \( P_e(E) \) and \( P_e(U) \) are, respectively, the probabilities of error for the eavesdropper and the legitimate user, and \( H(x) \) is the binary entropy function and follows,

\[
H(x) = -x log_2(x) - (1 - x) log_2(1 - x) \tag{6.12}
\]
6.4.2 Results and Discussion

The simulation and experimental results are discussed to validate the performance of WiLP, as shown below. A full chain is implemented on MATLAB based on the WLAN toolbox. The simulated BER performance for 16-QAM with coding rate = 1/2 is shown in Fig. 6.8A, where the BER curve of WiLP is slightly outperforming the overlapping RF and optical BER curves. The WiLP packets are partially transmitted on both the RF and optical links with the real part suffering from the limited indoor multi-path RF fading channel impact while the imaginary part is impacted by an optical quasi-static channel. Figure 6.8A also shows the eavesdropper’s performance that is above $10^{-1}$ BER even for high SNR values up to 15dB. For WiLP, the BER performance using different modulation schemes, i.e. MCS values from Table 17-4 in the IEEE 802.11 standards [98], is shown in Fig. 6.8B. The SNR requirements increase as the modulation order increases following the performance of a conventional QAM system.

The experimentally received RF constellations in Fig. 6.9A and the optical constellations in Fig. 6.9B demonstrate close performance. The experiment is implemented within a lab environment with limited RF interference sources at 85MHz. The one-dimensional constellations in Figs. 6.9C and 6.9D represent the partially received RF and optical signals, respectively. If either of these signals is received on its own, the receiver will not be able to estimate the transmitted bits. However, a legitimate user, can combine both constellations and obtain the WiLP constellations as in Fig. 6.9E.

The secrecy performance of the eavesdropper is measured by the secrecy capacity, as shown in (6.11). Using different MCS values in [98], the analytical normalized secrecy capacity is plotted against the SNR as shown in Fig. 6.10. As the secrecy capacity is approaching one, the system is more secure and the eavesdropper can not capture the data. The secrecy capacity is governed by the BER performance shown in Fig. 6.8B. For instance, at 4dB, 10dB and 17dB the secrecy capacity for MCS = 2, 4 and 6, in order, is at its maximum. This can be explained according to Fig. 6.8A, where below 4dB the BER performance of both the eavesdropper and WiLP is comparable.
Figure 6.8: (a) BER performance for WiLP using MCS = 4, i.e. 16-QAM with coding rate = 1/2 [98]. The performance is benchmarked against conventional RF and optical links. The eavesdropper’s BER is also shown, (b) BER performance for WiLP using different MCS orders.

Despite the bad performance for both, i.e. BER > 10^{-1}. Beyond 4dB, the eavesdropper’s BER performance saturates, while WiLP BER is enhanced with increasing SNR. Thus, the secrecy capacity starts to saturate beyond 4dB for QPSK, i.e. MCS = 2 [98]. The same applies for higher modulation orders as shown in Fig. 6.10 where the secrecy capacity saturates beyond 10 and 17dB for MCS = 4 and 6, respectively. The trade-off between the secrecy capacity and the system capacity can be optimized to obtain the maximum capacity and security at the target BER threshold.

The experimental setup is used to validate WiLP, i.e. the legitimate user’s BER against the eavesdropper’s BER, as shown in Fig. 6.11. Similar to the simulation results, the BER of the eavesdropper is above 10^{-1}, while the legitimate user can decode the transmitted data successfully. The experimental results validate the potential of the proposed WiLP design in terms of BER performance and secrecy capacity.
Figure 6.9: Experimentally received constellations for (a) RF link, (b) optical link, (c) partial RF link, (d) partial optical link, and (e) WiLP.
FIGURE 6.10: Secrecy capacity against SNR using different modulation schemes.

FIGURE 6.11: Experimentally Fitted and Simulated WiLP BER performance using 16-QAM with coding rate = 3/4 against eavesdropper’s BER.
6.5 Summary

This chapter proposes a novel perspective for new concepts of coordinated light sources and assistive links. The introduced concepts LwO-OFDM and WiLP serve low-complexity IoT receivers and PHY security, respectively. LwO-OFDM is compatible with indoor VLC localization and sensing beaconing systems. The study investigates the aggregation of multi-source single carriers; used mainly for positioning and illumination, in the optical domain to replace complex computation blocks, for example FFT and thus, enable simple low-cost IoT devices to receive and detect OFDM coded bits through FFT-less OFDM receivers. The findings of this investigation support the proposed scheme for its enhanced BER performance and low PAPR compared to conventional single source OFDM techniques. In addition, the results implicate the same spectral efficiency as in conventional ACO-OFDM while reducing the LEDs spectrum requirements.

To the authors’ best knowledge, this is the first time to introduce one possible realization of the WiLP approach. Such a novel PHY security feature is enabled through a joint RF-optical signal design in the digital-domain and simultaneous RF-optical transmission. The design is based on the IEEE 802.11 standards for WLAN networks and demonstrated in hybrid RF-optical communications. The obtained results highlight the potential of WiLP in delivering an effective BER performance slightly better than a standalone RF or optical link while simultaneously securing the transmission against an eavesdropper’s attack. BER and secrecy capacity performances are validated by simulation and experimental results.
Chapter 7

Conclusions and Future Work

7.1 Conclusions

This thesis introduces, for the first time, novel PHY concepts and designs to support the 6G vision of futuristic wireless networks. The proposed designs support

- heterogeneity of devices, interfaces, technologies and services.

- power efficiency requirements and battery life extension for IoT devices.

- security, privacy and secrecy of wireless networks as a key challenge in designs of futuristic modern networks.

In order to support these requirements, this thesis proposes

- **MCC** that is considered the first device-independent communication PHY design that provides concurrent wireless services including broadband access, low-rate IoT connectivity, device-free sensing, and device-based localization. System optimization to study system design parameters is implemented. Simulation and experimental investigation proves the validity of the concept and design to support high and low speed communication concurrently with positioning beacons while maintaining illumination control. Proper design promotes immunity to inter-service interference. The obtained results show that an optimum SNR threshold of 25.1dB is recommended for 64-QAM to avoid clipping distortion. A desired
BER performance below $10^{-3}$ for SNR values above 18dB was presented. There is a trade-off between the SNR of the OFDM stream and the useful dimming range. However, a 54% dimming range was obtained at 24dB SNR. An interference-immune performance of a properly managed spectrum with an SNR penalty of 2-5dB is presented for different interference scenarios.

- **Futuristic concepts on assistive links**

  - **LwO-OFDM** which is a novel perspective of integrating multi-carrier modulation with IoT applications based on spatial summation techniques. The study explores the energy saving aspect for IoT devices by introducing a concept that replaces complex computation blocks, for example FFT. Additionally, it supports positioning and communication on top of illumination. The obtained results show better BER performance and lower PAPR compared to conventional OFDM systems. In addition, the results implicate the same spectral efficiency as in conventional ACO-OFDM while reducing the LEDs spectrum requirements.

  - **WiLP** is introduced as a PHY security technique based on hybrid RF and optical links. The system design is highlighted based on IEEE 802.11 standards for WLAN networks. Early stage simulation and experimental results will are investigated and discussed to show immunity to eavesdropping at a reliable BER performance.

### 7.2 Future Work

In this section, the research plan for the development of the concepts introduced within this thesis is detailed.
7.2.1 Future of MCC Research

The novelty of MCC concept and design opens up varying research topics. MCC is introduced for VLC in this thesis. Further investigation of introducing the MCC concept for RF is essential. This study will enable a possible design that can maintain the same functionality within a challenging RF environment with its own wireless standards to support the 6G vision of technology heterogeneity. Moreover, different interference sources are to be modelled to explore the impact of surrounding environment on the overall MCC performance and per individual service in a realistic room model scenario.

7.2.2 LwO-OFDM Research Plan

The potential of the introduced concept and design requires more experimental work and validation. The next milestones list the research steps towards integrating this concept under the umbrella of MCC for 6G networks.

- Experimental investigation of the proposed approach is currently ongoing with test bench equipment. In addition to validating the concept and design, this will explore the gained power efficiency which directs towards the requirements of 6G of more power savings.

- Investigating the three-dimensional (3D) room model of the introduced concept will lead to proper system optimization. For instance, the number of sources required for proper illumination while maintaining a high quality of communication can be obtained, as well as, proper clustering of sources to serve specific locations. Additionally, the role of localization and communication as a top tier above the MCC design to be explored.

- Introducing the RF version of lightweight channel estimation and sensing.
7.2.3 WiLP Future Plan

Hybrid wireless networks are foreseen to play a major role in the visioning and planning of 6G communications. Most of the 6G applications are human-centric, thus, high security and privacy are key features. Therefore, the introduced concept is extensively investigated from a signal processing perspective to enable a design that can build on top of MCC with enhanced security performance. There is current work that has promising preliminary results related to this concept.

- Introducing different WiLP versions based on current limitations of both RF and VLC technologies is the first step. The introduced designs should maintain the bigger vision of MCC. WiLP should be investigated within the full picture of heterogeneity applications including, traffic offloading, capacity aggregation and technology coexistence.

- Adaptive algorithms should be implemented to support different WiLP modes within varying environments and conditions.
Appendix A

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